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Switched Reluctance Generators and Their Control

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9.1 Introduction

Switched reluctance generators (SRGs) are double-saliency electric machines with nonoverlapping stator multiphase windings and with passive rotors. They may also be assimilated with stepper motors with position-controlled pulsed currents. Multiphase configurations are required for smooth power delivery and eventual self-starting and motoring, if the application requires it.

SRGs were investigated mainly for variable speed operation as starter/generators on hybrid electric vehicles, as power generators, on aircraft and for wind energy conversion. They may also be considered for super-high-speed gas turbine generators from kilowatt to megawatt (MW) power range per unit.

As SRGs lack permanent magnets (PMs) or rotor windings, they are low cost, easy to manufacture, and can operate at high speeds and in high-temperature environments.

In vehicular applications, an SRG is required to perform over a wide speed range to comply with the internal combustion engine (ICE) that drives it. For wind energy conversion, limited speed range is needed to extract additional wind energy at lower mechanical stress in the system.

Aware of the very rich literature on SRMs [1, 2], we will treat in this chapter the following aspects deemed as representative:

- Practical topologies and principles of operation
- Characteristics for performance evaluation
- Design for wide constant power range
- Converters for SRG motor (M)
- Control of SRG as starter/generator with and without motion sensors

The existence of a handful of companies that fabricate and dispatch SRMs [3] and vigorous recent proposals of SRGs as starters/alternators for automobiles and aircraft (up to 250 kW per unit) seem sufficient reason to pursue the SRG study within a separate chapter such as this one.

9.2 Practical Topologies and Principles of Operation

A primitive single-phase SRG(M) configuration with two stator and two rotor poles is shown in Figure 9.1a and Figure 9.1b. It illustrates the principle of reluctance machine, where torque is produced through magnetic anisotropy. The stored magnetic energy (W_e) or coenergy (W_c) varies with rotor position to produce torque:

$$T_e = \left(\frac{\partial W_c(i, \theta_r)}{\partial \theta_r} \right)_{i=\text{cons}} = - \left(\frac{\partial W_c(\Psi, \theta_r)}{\partial \theta_r} \right)_{\Psi=\text{cons}} \quad (9.1)$$

$$W_c = \int_0^i \Psi di; \quad W_e = \int_0^\Psi i d\Psi \quad (9.2)$$

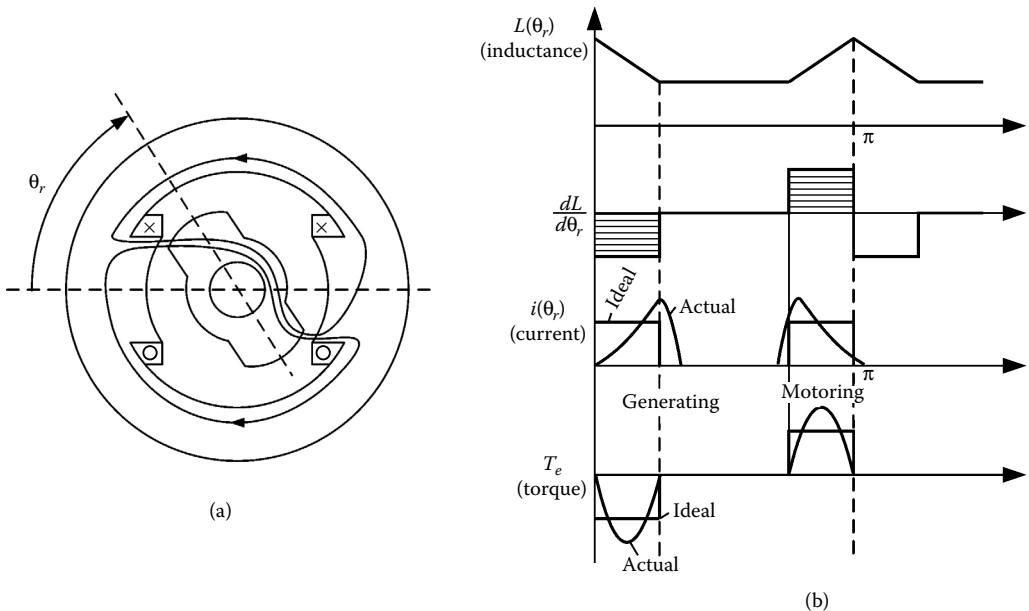


FIGURE 9.1 Primitive switched reluctance generator (SRG) (M) with (a) two stator and rotor poles and (b) ideal waveforms.

The energy conversion ratio (ECR) is as follows:

$$ECR = \frac{W_{mec}}{W_{mec} + W_{mag}} \geq 0.5 \quad (9.7)$$

For a SRG with the same maximum flux and peak current but larger airgap, when unsaturated, the ECR is around 0.5. For smaller airgap and saturated SRGs, it is larger (Figure 9.2).

SRGs require notably smaller airgaps than PM machines to reach magnetic saturation at small currents. The same small airgap, however, leads to notable vibration and noise problems, due to large local radial forces.

Three- and four-phase configurations have become commercial for SRM drives due to their self-starting capability from any rotor position and torque (power) pulsation reduction opportunities through adequate current/position profiling. The basic 6/4, 8/6 three-phase and, respectively, four-phase topologies are shown in Figure 9.3a and Figure 9.3b.

Ideal phase inductances vs. rotor position for the three- and four-phase machines are shown, respectively, in Figure 9.4a and Figure 9.4b. It should be noticed that with the three-phase machines, only one phase produces positive (or negative) torque at a time ($dL/d\theta_r \neq 0$), while two phases are active at all times in the four-phase machine. Low torque pulsations through adequate current waveform control with phase torque sharing are, thus, more feasible with four phases. To increase the frequency of the pulsations, the number of rotor and stator poles should be increased.

However, the energy conversion tends to deteriorate above a certain number of poles, for given rotor (stator) outer diameter, due to flux fringing and increased rotor core losses. In general, three or four phases are used, but the number of stator and rotor poles may be increased so as to have more such units per stator periphery:

$$\begin{aligned} N_s/N_r &= 6/4, 12/8, 18/12, 24/16 \dots \\ N_s/N_r &= 8/6, 16/12, 24/18, \dots, 32/24 \end{aligned} \quad (9.8)$$

An even number of stator sections (pole pairs) is appropriate when dual-output (two-channel) SRG operation is required.

To reduce the interaction flux between the two sections, the sequence of phase pole polarities along the rotor periphery (for a three-phase 12/8 pole combination) should be N N S S and *not* N S N S

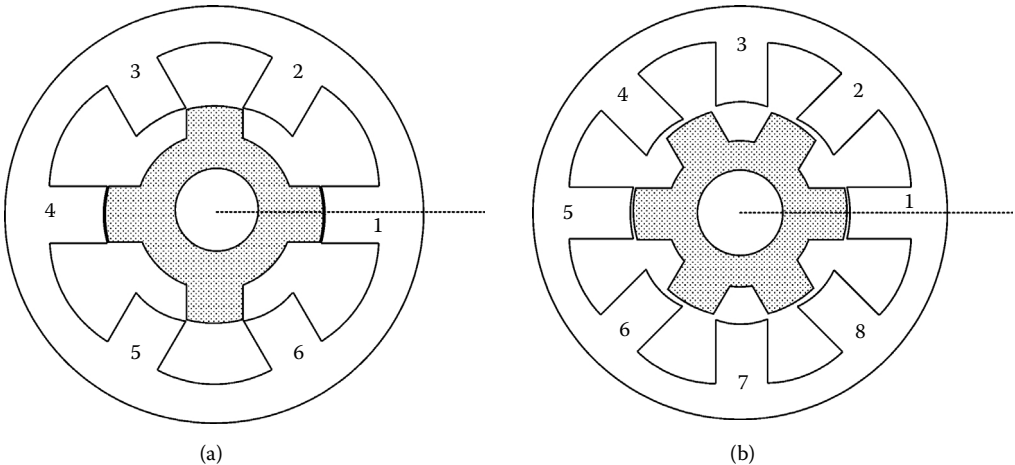


FIGURE 9.3 (a) Three- and (b) four-phase switched reluctance generators (SRGs) motors (Ms) with 6/4 and 8/6 stator/rotor pole combinations.

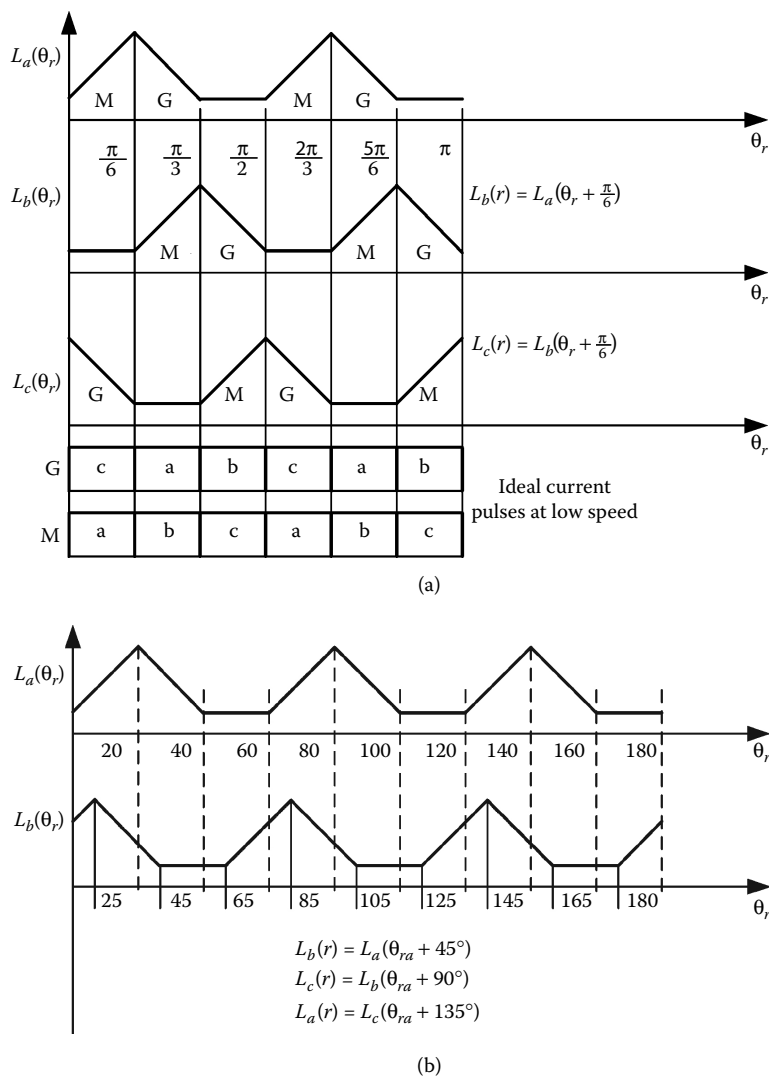


FIGURE 9.4 Ideal phase inductance/position dependence for the (a) three-phase and (b) four-phase switched reluctance generators (SRGs) motors (Ms).

standard sequence. However, in this case, the flux in the rotor and stator yokes of the two channels adds up, and thus, thicker yokes are required. The standard sequence per phase N S N S is less expensive, though more interference between channels is expected.

Returning to the principle of operation, we should note that for motoring, each phase should be connected when the phase inductance is minimal and constant, because, in this case, the emf is zero at any speed, and thus, the phase voltage equation reduces to the following:

$$V_{dc} = R \cdot i + L_u \frac{di}{dt} \tag{9.9}$$

Neglecting the phase resistance voltage drop, the flux accumulated in the phase, Ψ , at constant speed n (rpsec) is

$$\Psi = \int L_u \frac{di}{dt} dt = \int V_{dc} dt \approx V_{dc} \frac{\theta_w}{2\pi n} \tag{9.10}$$

Here, θ_w is the mechanical dwell (conduction) angle of the phase when one voltage pulse is applied. The maximum (ideal) value of θ_w is π/N_r .

In most designs, the value of the maximum flux Ψ_{max} is used to calculate the base speed of the SRG:

$$2\pi n_b = \frac{V_{dc} \theta_w}{\Psi_{max}} \tag{9.11}$$

θ_w is in mechanical radians. This is, in fact, equivalent to the ideal standard condition that the emf equals phase voltage for constant current and zero phase resistance.

The single voltage pulse operation, characteristic of high speeds, is illustrated in Figure 9.5a through Figure 9.5d (upper part). The low-speed operation appears in the lower part of Figure 9.5a (current

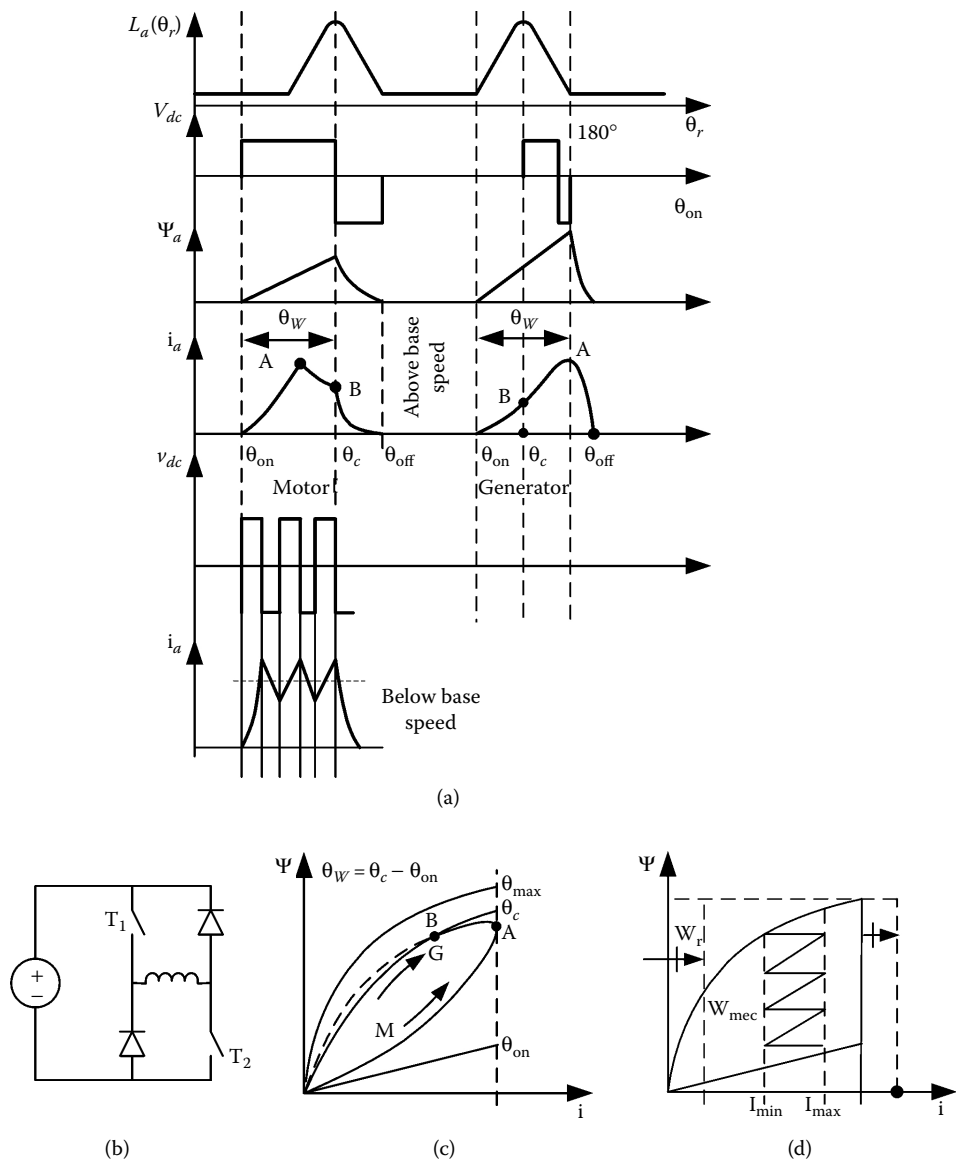


FIGURE 9.5 High-speed single voltage pulse and low-speed pulse-width modulator (PWM) voltage pulse operation: (a) waveforms, (b) phase converter, (c) single pulse energy cycle, and (d) energy cycle with PWM.

chopping). While current chopping is typical for motoring below base speed, generating is performed, in general, above base speed in the single voltage pulse mode. In special cases, to reduce regenerative braking torque, a pulse-width modulator (PWM) may also be used for the generator mode.

To vary the generator output, only the angles θ_{on} and θ_c may be varied, in general. The turn-on angle θ_{on} may be advanced at high speeds, both for motoring and generating, to produce more torque (or power). The negative voltage pulses refer to the so-called hard switching, where both controlled power switches T_1 and T_2 (Figure 9.5b) are turned off at the same time, and the free-wheeling diodes become active.

The energy cycle is traveled from A to B for motoring and from B to A for generating (Figure 9.5c). The PWMs of voltage effects are shown in Figure 9.5d.

Increasing the energy cycle area W_{mec} means increasing the torque. This is possible by adding a diametrical DC-fed coil to move the energy cycle to the right in Figure 9.5d. In essence, the machine is no longer totally defluxed after each energy cycle. Alternatively, the continuous current control in a phase would lead to similar results, though at the price of additional losses.

Besides torque density and losses, which refer essentially to machine size and goodness, the kilowatt to peak kilovoltampere (kW/peak kVA) ratio defines the ratings of the static converter needed to control the SRG(M).

9.2.1 The kW/Peak kVA Ratio

It was shown [1] that the kW/peak kVA ratio for SRG(M) is as follows:

$$kW/peak\ kVA \approx \frac{\alpha_s \cdot N_r \cdot Q}{8\pi} \quad (9.12)$$

where $\alpha_s = 0.4$ to 0.5 is the stator pole ratio and Q is as follows:

$$Q \approx C_{m,g} \left(2 - \frac{C_{m,g}}{C_s} \right) \quad (9.13)$$

C_m is the ratio between the active dwell angle θ_{wu} and the stator pole span angle β_s . $C_m = 1$ only at zero speed, and then it decreases with speed and reaches values of 0.6 to 0.7 at base speed.

For the generator mode,

$$C_g = \left(1 - \frac{\theta_{wu}}{\beta_s} \right) \quad (9.14)$$

where θ_{wu} is the angle from phase turnoff (after magnetization), when active power delivery starts. Again, $C_g = 0.6$ to 0.7 should be considered acceptable.

The coefficient C_s [1] is as follows:

$$C_s = \frac{\lambda_u - 1}{\lambda_u \cdot \sigma - 1}; \quad \lambda_u = \frac{L_u^u}{L_u} \approx 4-10; \quad \sigma = \frac{L_s^u}{L_a^u} = 0.25-0.4 \quad (9.15)$$

where

L_a^u is the aligned unsaturated inductance per phase

L_u^u is the unaligned inductance

L_s^u is the aligned saturated inductance

The peak power S of the switches in the SRG(M) converter is as follows:

$$S = 2 \cdot m_1 \cdot V_{dc} \cdot I_{peak} \quad (9.16)$$

For an inverter-fed IM (or alternating current [AC] machine),

$$kW/peak\ kVA \approx \frac{3 \cdot V_{dc} \cdot I_{peak} \times P.F.}{\pi \cdot K \cdot 6 \cdot V_{dc} \cdot I_{peak}} = \frac{3 \times P.F.}{\pi \cdot 6 \cdot K} \quad (9.17)$$

where K is the ratio between peak current waveform value and its fundamental peak value. For the six-pulse mode of the PWM converter $K = 1.1$ to 1.15 . $P.F.$ is the power factor for the fundamental.

Example 9.1

Consider a 6/4 three-phase SRG(M) with $\sigma = 0.3$, $\lambda_u = L'_a/L_a = 8$, $C_g = 0.8$, $\alpha_s = 0.45$, and calculate the kW/peak kVA ratio. Compare it with an induction machine (IM) drive with $P.F. = 0.81$ and $K = 1.12$. From Equation 9.15,

$$C_s = \frac{\lambda_u - 1}{\lambda_u \sigma - 1} = \frac{8 - 1}{8 \cdot 0.3 - 1} = 0.5$$

The value of Q comes from Equation 9.13:

$$Q \approx C_g \cdot \left(2 - \frac{C_g}{C_s} \right) = 0.8 \cdot \left(2 - \frac{0.8}{0.5} \right) = 1.472$$

Finally, from Equation 9.12,

$$(kW/peak\ kVA)_{RSG} = \frac{0.45 \cdot 4 \cdot 1.472}{8 \cdot \pi} \cong 0.1055$$

For the IM drive (Equation 9.17),

$$(kW/peak\ kVA)_{IM} = \frac{3 \cdot 0.85}{6 \cdot \pi \cdot 1.12} = 0.1208$$

For equal active power and efficiency, the IM requires from the converter about 10 to 15% less peak kVA rating.

When the cost of the converter per SRG cost is large, larger system costs with the SRG(M) are expected.

Note that the kW/peak kVA as defined in this example is not equivalent to $P.F.$, but it is a key design factor when the converter rating and costs are considered.

An equivalent $P.F.$ for SRG(M) may be defined as follows [4]:

$$(P.F.)_{SRM} = \frac{\text{output shaft power}}{\text{input r.m.s. volt} \cdot \text{ampere}}$$

For given power, this $P.F.$ varies with speed, and for the very best designs in Reference [4], it is above 0.7 and up to 0.86.

However, it is the peak value of current, not its RMS value, that determines the converter kVA rating.

From this point of view, AC drives seem slightly superior to SRG(M)s.

9.3 SRG(M) Modeling

It was proven through detailed finite element method (FEM) analysis that the interaction between phases in standard SRG is minimal. Consequently, the effects of various phases may be superposed:

$$V_{a,b,c,d} = R_s i_{a,b,c,d} + \frac{d\Psi_{a,b,c,d}(\theta_r, i_{a,b,c,d})}{dt} \quad (9.18)$$

A four-phase SRG(M) is considered here.

Only the family of flux/current/position curves for one phase is required, as periodicity with π/N_s exists. The $\Psi(\theta_r, i)$ curves may be obtained from experiments or through calculation via analytical methods or FEM. The torque per phase (Equation 9.1 and Equation 9.2) becomes

$$T_{e,a,b,c,d} = \frac{\partial}{\partial \theta_r} \int_0^{i_{a,b,c,d}} \Psi_{a,b,c,d}(\theta_r, i_{a,b,c,d}) di_{a,b,c,d} \quad (9.19)$$

The total torque is

$$T_e = \sum_{a,b,c,d} T_{e,a,b,c,d} \quad (9.20)$$

The motion equations are

$$J2\pi \frac{dn}{dt} = T_e - T_{load}; \quad \frac{d\theta_r}{dt} = 2\pi n \quad (9.21)$$

The phase i voltage Equation 9.18 may be written as follows:

$$V_i = R_s i_i + \frac{\partial \Psi_i}{\partial i} \frac{di_i}{dt} + \frac{\partial \Psi_i}{\partial \theta_r} \frac{d\theta_r}{dt} \quad (9.22)$$

The transient inductance L_{ii} is defined as

$$L_{ii} = \frac{\partial \Psi_i}{\partial i} = L_i(\theta_r, i_i) + i_i \frac{\partial L_i(\theta_r, i_i)}{\partial i_i}; \quad L_i = \frac{\Psi_i}{i_i} \quad (9.23)$$

The last term in Equation 9.22 represents a pseudo-emf E_i :

$$E_i = \frac{\partial \Psi_i}{\partial \theta_r} \cdot 2\pi n = K_E(\theta_r, i_i) \cdot 2\pi n \quad (9.24)$$

E_i is positive for motoring and negative for generating. V_i is considered positive and equal to V_{dc} when the DC source is connected through the active power switches to the SRG, it is zero when only one switch is turned off (soft commutation), and it is $(-V_{dc})$ when both active power switches are turned off (hard commutation).

In a well-designed SRG(M), the $\Psi(\theta_r, i)$ family of curves shows notable nonlinearity (Figure 9.2). Consequently, both the transient inductance L_i and the emf coefficient K_E are dependent on rotor position and on current [5].

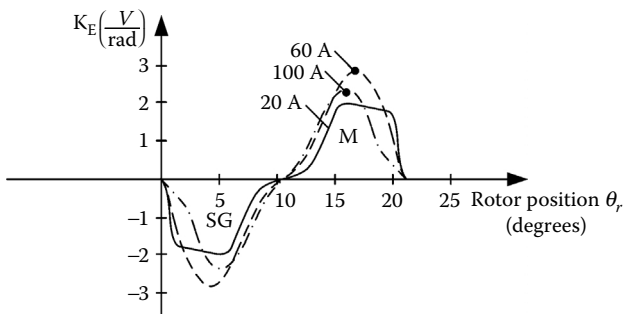


FIGURE 9.6 Electromagnetic force (emf) coefficient vs. rotor position for various currents.

Typical emf coefficients K_E , calculated through FEM, are shown qualitatively in Figure 9.6. It should be noted that K_E , for constant current, is notably variable with rotor position; it is positive for motoring and negative for generating.

The voltage equation suggests an equivalent circuit with variable parameters (Figure 9.7a and Figure 9.7b). The source voltage V_i is considered positive during phase energization and negative or zero during phase de-energization.

It should be emphasized that the emf E is, in fact, a pseudo-emf (when the machine is magnetically saturated), as it contains a small part related to stored magnetic energy. Consequently, the instantaneous electromagnetic torque has to be calculated only from the coenergy (Equation 9.17), for a magnetically saturated machine [6].

Iron loss occurs in both the stator and the rotor, and this loss is due to current vs. time variation and to motion at rectangular current in the phases [2]. So, in the equivalent circuit, we should “hang” core resistances R_{cm} and R_{ct} in parallel to the transient inductance voltage and around the pseudo-emf (Figure 9.7). R_{cm} and R_{ct} reflect the core loss presence in the model.

9.4 The Flux/Current/Position Curves

For refined design attempts and for digital system simulations of SRG for transient and control, the nonlinear flux/current/position family of curves has to be put in some analytical, easy-to-handle form. Even simpler expressions are required for control implementation.

Through FEM calculations, the family of such curves is obtained first, and then curve fitting is applied. The problem is that it is not enough to curve-fit $\Psi(\theta_r, i)$, but to determine also $i(\Psi, \theta_r)$ and, eventually, $\theta_r(\Psi, i)$. Polynomial or exponential functionals, fuzzy logic, artificial neural network (ANN) or other methods of curve fitting were proposed for this function [2].

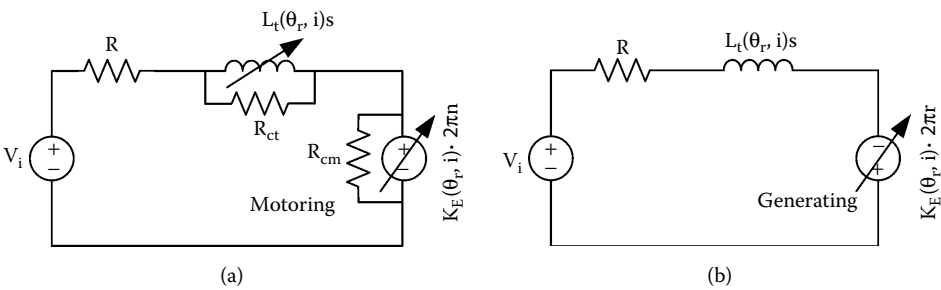


FIGURE 9.7 The equivalent circuit: (a) for motoring and (b) for generating.

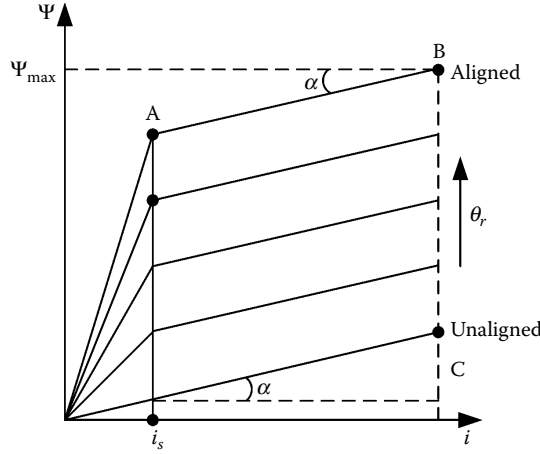


FIGURE 9.8 Piecewise linear approximations of the magnetization curves.

Examples of exponential approximations are as follows [7]:

$$\Psi(\theta_r, i) = a_1 \cdot (\theta_r) \cdot (1 - e^{-a_2 \cdot (\theta_r) \cdot i}) + a_3 \cdot (\theta_r) \cdot i \quad (9.25)$$

with

$$a_{1,2,3}^{(\theta_r)} = \sum_{K=0}^{\infty} A_{1,2,3}^K \cos(K \cdot N_r \cdot \theta_r) \quad (9.26)$$

or

$$\Psi_j = \Psi_{sat} (1 - e^{-i_j f_j(\theta_r)}); \quad i_j \geq 0 \quad (9.27)$$

$$f_j(\theta_r) = a_0 + \sum_{k=1}^3 a_n \cdot \cos\left(k \cdot N_r \cdot \theta_r - (j-1) \frac{2\pi}{N_s}\right); \quad j=1,2,3,(4) \quad (9.28)$$

From Equation 9.27, the inverse function $i_j(\Psi_j, \theta_r)$ is as follows:

$$i_j = \frac{1}{f_j(\theta_r)} \ln\left(\frac{\Psi_s}{\Psi_s - \Psi_j}\right) \quad (9.29)$$

Ψ_s is the saturated value of the flux for the aligned position.

The coenergy (Equation 9.17) with Equation 9.27 becomes

$$W_{cj} = \int_0^i \Psi_j di_j = \int_0^i \Psi_{sat} (1 - e^{-i_j f_j(\theta_r)}) di_j = \Psi_{sat} \left[i_j - \frac{(1 - e^{-i_j f_j(\theta_r)})}{f_j(\theta_r)} \right] \quad (9.30)$$

So, the instantaneous torque per phase is

$$T_{ej} = \frac{\partial W_{cj}}{\partial \theta_r} = - \Psi_{sat} \frac{\partial}{\partial \theta_r} \left[\frac{(1 - e^{-i_j f_j(\theta_r)})}{f_j(\theta_r)} \right] \quad (9.31)$$

or

$$T_{ej} = \frac{-(\partial f_j / \partial \theta_r)}{f_j^2} (\Psi_{sat} i_j f_j + \Psi_j (1 - i_j f_j)) \quad (9.32)$$

For control design purpose, even a piecewise linear approximation of magnetization curves may be acceptable, as test results confirm that, in practical designs, the magnetic saturation curve corner occurs at about the same current, independent of rotor position (Figure 9.8). This, in fact, implies that right after some stator/rotor pole overlapping, local saturation is achieved, and thus, the flux varies almost linearly with rotor position.

Consequently,

$$\begin{aligned} \Psi_j &= i_j \left(L_u + \frac{K_s (\theta_r - \theta_0)}{i_s \beta_s} \right) \quad \text{for } i \leq i_s \\ \Psi_j &= i_j L_u + \frac{K_s (\theta_r - \theta_0)}{\beta_s} \quad \text{for } i \geq i_s \end{aligned} \quad (9.33)$$

The only variables to be found from the family of magnetization curves are i_s and K_s . The angle θ_0 corresponds to the rotor position where the rotor poles start overlapping the stator poles of the respective phase. While this simple approximation is tempting, because the continuity of flux at i_s is maintained, the continuity of the $\partial \Psi / \partial i$ at i_s is not preserved. However, $\partial \Psi / \partial \theta_r$ is continuous at $i = i_s$. Knowing only the positions of A, B, and C on the graph in Figure 9.8 leads to unique values of K_s , i_s , and L_u . It suffices to calculate only the fully aligned and unaligned position magnetization curves for this approximation. A thorough analytical model, with FEM verifications, is developed in Reference [8].

9.5 Design Issues

The comprehensive design of SRG(M)s is complex. A few basic issues are listed here:

- Motor and generator specifications
- Number of stator phases m and stator and rotor poles N_s, N_r
- Stator bore diameter D_{is} and stack length l_{stack} calculation
- Computation of stator and rotor pole and yoke geometry
- Number of turns per coil, conductor gauge, and connections
- Loss computation model
- Thermal model and temperature vs. speed for specific duty-cycle operation modes
- Magnetization curve family and curve fitting for the design of the control system
- Peak and rated current, torque, and losses vs. speed

Some of the above subjects will be detailed in what follows.

9.5.1 Motor and Generator Specifications

The design specifications are tied to the application. For a generator-only application (wind generator, auxiliary power generator on aircraft, etc.), the motoring mode is excluded from start, while the DC output voltage power bus may be as follows:

- Independent, with passive and active loads
- With a battery backup

The DC voltage may be constant for stand-alone operation but variable when, say, connected to a grid through an additional PWM inverter interface. For wind generation, a $\pm 30\%$ variation of speed around base speed is sufficient for most practical situations.

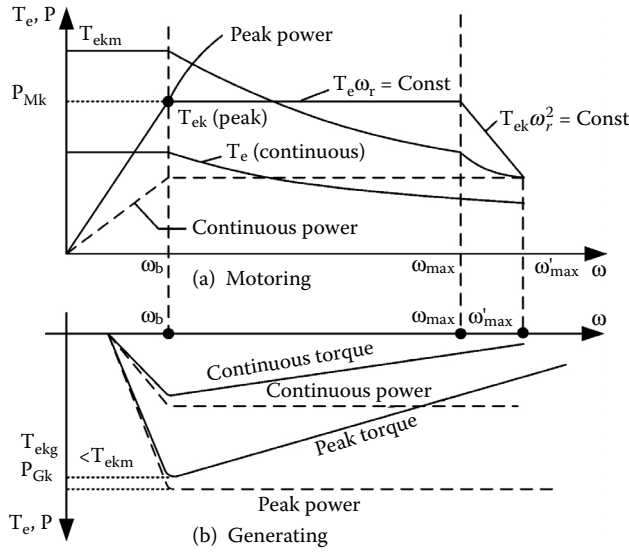


FIGURE 9.9 Typical torque and power vs. speed requirements for starter generators: (a) motoring and (b) generating.

Also, for the backup battery load case, the voltage varies from a minimum to a maximum value $V_{dc} = V_{dc}^r(1 \pm (0.2 - 0.25))$. The battery backup operation stabilizes the control requirements on the generator.

For automotive starter/generators both motoring and generating are mandatory. The constant power speed range for motoring is generally larger than 3–4 to 1, but for generating (at lower power), much larger values are welcome, and a 10:1 ratio is not unusual (Figure 9.9).

The design DC voltage V_{dc} is very important when calculating the number of turns per winding and, consequently, the peak current for peak torque. A boost DC–DC converter might be added to raise the voltage to the battery maximum voltage and to reduce the peak machine and PWM converter current.

9.5.2 Number of Phases, Stator and Rotor Poles: m , N_s , N_r

The fundamental frequency in each phase f_1 is as follows:

$$f_1 = n \times N_r; \quad n - \text{speed in rps} \quad (9.34)$$

The number of strokes per second f_s is

$$f_s = m \cdot f_1; \quad m - \text{number of phases} \quad (9.35)$$

The number of stator poles per phase is a multiple of two (pole pairs):

$$N_s = m \cdot 2K; \quad K - \text{integer} \quad (9.36)$$

The number of stator and rotor poles are generally related by

$$N_s - N_r = 2K \quad (9.37)$$

though other combinations are feasible.

Due to the increased complexity and costs in the PWM converter, only three- and four-phase SRG(M)s are considered practical for most applications above 1 kW power per unit.

Increasing the number of poles N_s , for given outer stator diameter and stack length, leads to a reduction in the stator and rotor yoke thickness, without leading to higher torque to the same extent.

The frequency f_1 increases and so does f_s , which determines the core losses that tend to increase.

On the other hand, the noise level tends to decrease notably as the radial force of the active phase is distributed around the rotor periphery more evenly.

It was established that, in general, the stator and rotor pole heights h_{ps} , h_{pr} per pole spans, b_{ps} , b_{pr} should be as follows:

$$\frac{h_{ps}}{b_{ps}} \approx \frac{h_{pr}}{b_{pr}} \approx (0.7 - 0.8) \quad (9.38)$$

Also, the pole span per airgap g should be

$$\frac{b_{ps}}{g} \approx \frac{b_{pr}}{g} > (17 - 20) \quad (9.39)$$

to yield enough high ratios between aligned and unaligned maximum inductance ($L_a^s/L_u \approx 3 - 8$) to guarantee acceptably high average torque (W_{mec}).

The airgap should be as small as mechanically feasible, that is, for acceptable vibration deflection and noise levels.

The maximum frequency f_1 is also related to the switching capacity of the PWM converter, as the commutation converter losses (Chapter 8) increase with frequency.

9.5.3 Stator Bore Diameter D_{is} and Stack Length

Though SRG(M)s operate at variable speed, the Epson's coefficient heritage may be used to calculate the stator bore diameter D_{is} , assuming first that $\lambda = 1$ stack/ b_{ps} and the number of stator poles N_s are given. Alternatively, the shear rotor stress f_t may be imposed:

$$f_t = (1 + 10)N/cm^2 \quad (9.40)$$

$$\lambda = \frac{l_{stack}}{b_{ps}} \approx 0.5 - 2.0; \quad b_{ps} \approx \frac{\pi D_{is}}{2N_s}$$

The design tangential peak shear rotor stress may increase with stator bore diameter. Once the peak torque T_{ek} is given through the specifications, the stator bore diameter D_{is} is as follows:

$$D_{is} = \sqrt[3]{\frac{4 \cdot N_s \cdot T_{ek}}{\pi^2 \cdot f_{tk} \cdot \lambda}} \quad (9.41)$$

Observing Equation 9.38 and Equation 9.39, we still need to size the stator and rotor yokes to complete the stator geometry. In general, h_{ys} $h_{yr} > b_{ps}/2$, as half the stator pole flux flows through the yokes. The peak torque is needed for motoring at low speeds, when the current is kept constant through chopping. Consequently, the energy cycle area is full. Making use of FEM or of an analytical model, the unaligned and aligned position flux curves are obtained. Φ_{pole} is the flux per stator pole for active phase, and $W_c I_c$ is the corresponding coil magnetomotive force (mmf; Figure 9.10). The maximum flux per pole is related to the saturation flux density and the pole area, plus the leakage flux (by K_l):

$$\Phi_{pmax} = B_{sat} \cdot b_{ps} \cdot l_{stack} \cdot (1 + K_l) \quad (9.42)$$

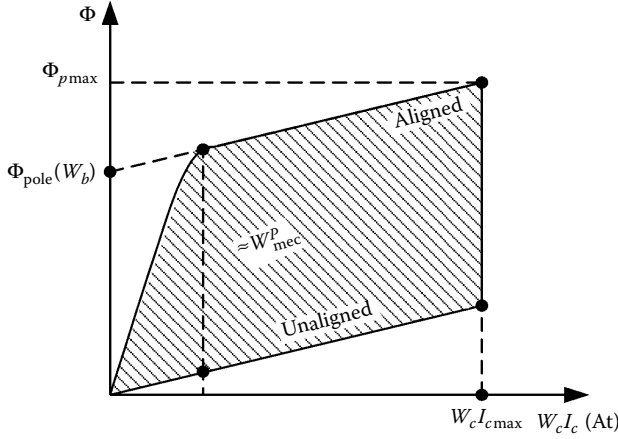


FIGURE 9.10 Extreme position magnetization curves.

With $B_{sat} = 1.6$ to 2 T, b_{ps} , I_{stack} known already, and $K_1 \approx 0.1$ to 0.2 , the area W_{mec}^p in Figure 9.10 leads to the average peak torque:

$$T_{ek} = \frac{W_{mec}^p \cdot m \cdot N_r \times 2K}{2\pi} = \frac{W_{mec}^p \cdot N_s \cdot N_r}{2\pi} \quad (9.43)$$

with $2K$ equal to the number of stator poles per phase. As T_{ek} is fixed by specifications, the value of B_{sat} may be increased gradually until W_{mec} is high enough to satisfy Equation 9.43, for a certain $W_c I_{cmax}$.

The window area available for a coil A_{Wc} is as follows:

$$A_{Wc} \approx \left[\frac{\pi}{4} \left((D_{is} + 2h_{ps})^2 - D_{is}^2 \right) - N_s \cdot b_{ps} \cdot h_{ps} \right] / 2N_s \quad (9.44)$$

For a window filling $K_{fill} = 0.5$ to 0.6 , typical for preformed coils, the maximum current density j_{COmax} is obtained:

$$j_{COmax} = \frac{W I_{cmax}}{A_{Wc} K_{fill}} \quad (9.45)$$

Depending on the application, j_{COmax} may vary from 8 – 10 A/mm² to 30 – 40 A/mm² for most strained automotive starter/generators with forced cooling. If j_{COmax} is out of this range, the design is restarted with a lower shear rotor stress f_{tk} .

9.5.4 The Number of Turns per Coil W_c for Motoring

The $2K$ coils per phase may be connected many ways to form from 1 to $2K$ current paths in unity steps $a = 1, 2, \dots$. The number of turns in series per current path W_a is related to the total number of turns per phase by the following:

$$W_a = \frac{W_c \cdot N_s / m}{a} \quad (9.46)$$

The machine is designed so that the ideal emf E_{av} , for constant current and full angle conduction,

$$\theta_w = \theta_c - \theta_{on} \approx \frac{2b_{ps}}{D_{is}} \quad (9.47)$$

at base speed n_b (rpsec), be a given fraction α_{PF} of the DC design voltage V_{dc} (Equation 9.11):

$$\frac{2\pi \cdot n_b \cdot (\Psi_{\max})_{\text{path}}}{\theta_w} = \frac{2\pi \cdot n_b}{\theta_w} \cdot \Phi_{p\max} \cdot W_a = \alpha_{PF} \cdot V_{dc} = E_{av} \quad (9.48)$$

From this expression, with $\Phi_{p\max}$ already calculated from Figure 9.10 and θ_w from Equation 9.47 and assigned values of n_b and α_{PF} , the number of turns W_a per current path is obtained. Subsequently, from Equation 9.46, the number of turns per coil W_c is calculated, and, with $W_c I_{c\max}$ known from Figure 9.10 for peak torque (Equation 9.43), the maximum coil current I_{peak} is as follows:

$$I_{\text{peak}} = a \cdot I_{c\max} \quad (9.49)$$

The coefficient $\alpha_{PF} = E_{av}/V_{dc}$ at base speed is a measure of the machine power factor, as in AC machines. The α_{PF} best practical values gravitate around unity. When a large constant power speed range is required, $\alpha_{PF} < 1$, but then the number of turns W_a is smaller. Consequently, the peak current is larger, and inverter oversizing is required. In contrast, with $\alpha_{PF} > 1$, the constant power speed range decreases, but W_a increases.

It seems wise to start with $\alpha_{PF} = 1$ and then investigate the performance for small departures from this situation, hunting for the best overall performance for the constant power speed range.

Randomly modifying W_c would eventually lead to similar results, but the search may be too time consuming. At this point, preliminary machine sizing is complete, and all parameters may be calculated. The average torque/speed envelopes for motoring and generating can be calculated up to the peak current and to the peak voltage (V_{dc}). This way, the generator specifications may be checked after the machine is sized in terms of turns/coil W_c , for motoring.

In some applications, the generator mode is more demanding; thus, the W_c computation should be tied to generating. To do so, we first have to explore current waveforms for generator mode.

9.5.5 Current Waveforms for Generator Mode

In the generator mode, a phase is turned on at the angle θ_{on} around the maximum inductance rotor position, and then the controlled semiconductor rectifier (SCR) is turned off (commutated) at the angle θ_c , long before the phase inductance decreases to its minimum value (Figure 9.11a through Figure 9.11c).

The voltage equation for phase j (Equation 9.22) is as follows:

$$V_j = R_j I_j + L_{tj} \cdot \frac{di_j}{dt} + K_E \cdot 2\pi \cdot n; \quad K_E = i \frac{dL}{d\theta} < 0 \quad (9.50)$$

For simplicity, we may consider $L_t \approx L_u$ (unaligned) = const, as it is the situation in very saturated conditions. The voltage $V_j = V_{dc}$ during phase energization (excitation) and $V_j = -V_{dc}$ during power delivery through the free-wheeling diodes (Figure 9.11a). Let us also neglect the stator resistance.

Figure 9.11a shows three cases that can be fully documented via Equation 9.50. In case (1), after T_1 and T_2 are turned off (phase energization is finished) at θ_c , the current still increases. As $V_j < 0$ and $E < 0$, the current increase is due to the fact that $|E| > |V_{dc}|$, that is, the emf is greater than the supply (magnetization) voltage. This case is typical for high speeds ($n > n_b$), when the torque is smaller (Figure 9.11b, cycle 1 area).

For case (2) in Figure 9.11a, it happens that at turn-off angle θ_c , $|E| = |V_{dc}|$ and thus, from Equation 9.50, with $R_s = 0$, $\frac{di}{dt} = 0$. Consequently, the current remains constant until the inductance reaches its minimum at θ_d . Energy cycle (2) in Figure 9.11b shows the large torque area.

In case (3) in Figure 9.11a, the maximum current is reached at θ_c , and after that, the current decreases steadily, because $|E| < |V_{dc}|$, which corresponds to low speeds. A smaller torque area is typical for case (3).

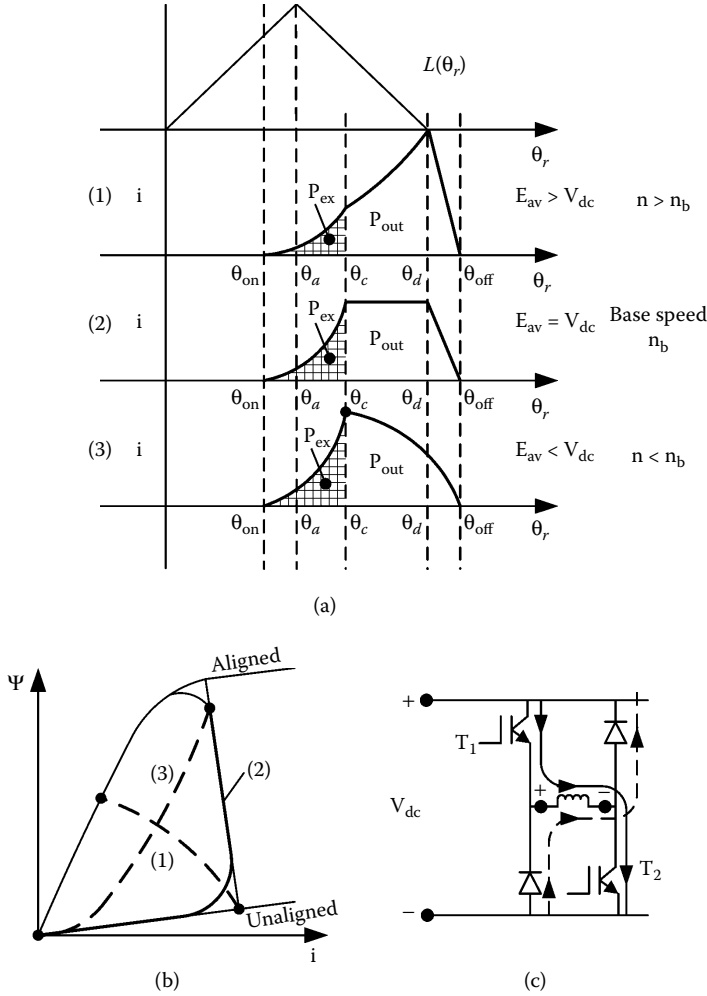


FIGURE 9.11 Current waveforms for (a) generator mode, (b) the corresponding energy cycles, and (c) the same peak current with pertinent converter.

For constant DC voltage power delivery, the excitation energy W_{exc} per cycle (Figure 9.11a) is as follows:

$$W_{exc} = \frac{V_{dc}}{2\pi n} \int_{\theta_{on}}^{\theta_c} i d\theta \quad (9.51)$$

The delivered ideal energy/cycle, W_{out} , is

$$W_{out} = \frac{V_{dc}}{2\pi n} \int_{\theta_c}^{\theta_{off}} i d\theta \quad (9.52)$$

The excitation penalty ε is

$$\varepsilon = \frac{W_{exc}}{W_{out}} \quad (9.53)$$

Case (1) ($|E| > |V_{dc}|$) leads to a smaller excitation penalty than that for the other two cases, for the same net generated energy per cycle W_{out} . As we can see from Figure 9.11a and Figure 9.11b, the excitation penalty has conflicting influences on the design. To have a small ε , we are tempted to design (control) the machine such that $|E| > |V_{dc}|$; however, the torque area (Figure 9.11b) is smaller, and the energy conversion ratio is not close to an optimum.

Case (2) ($E_{av} = V_{dc}$) produces the largest torque–energy cycle and, thus, seems more effective in energy conversion. Maintaining case (2) ($|E| = |V_{dc}|$) forces one to increase the DC voltage in proportion to speed. But, to deliver power at constant voltage, a voltage step-down DC–DC converter has to be added between the machine converter and the DC load (from V_{dc} to V_L). The peak value of current for case (1) occurs at the angle θ_d :

$$i_{peak} \approx \frac{\Psi_{peak}}{L_u} = \frac{1}{L_u} \frac{V_{dc}}{2\pi n} (\theta_{off} - \theta_d) \approx \frac{1}{L_u} \frac{V_{dc}}{2\pi n} (\theta_c - \theta_{on} + \theta_c - \theta_d) \quad (9.54)$$

An earlier turnoff (smaller θ_c) leads to a smaller peak current, even if $\theta_c - \theta_{on} = \text{const.}$

For case (2), the peak current occurs at θ_c :

$$i_{peak} = \frac{\Psi_{peak}}{L_{peak}} = \frac{1}{L} \cdot \frac{V_{dc}}{2\pi n} \cdot \frac{\theta_c - \theta_{on}}{\theta_d - \theta_c} \quad (9.55)$$

The phase inductance L_{peak} at $\theta_{peak} = \theta_c$ is considered to be proportional to $\theta_d - \theta_c$, because L varies almost linearly with rotor position. A constant dwell angle $\theta_c - \theta_{on}$ with early turnoff (smaller θ_c) leads to smaller peak current and, thus, lower power delivery. Controlling the output power is thus possible by turn-on angle θ_{on} and turn-off angle θ_c .

When a voltage step-down DC–DC converter is not available or desirable, the generator mode should switch from case (3) below base speed, to case (2), and case (1) as speed increases. The energy conversion is not uniformly good, but the constant power speed range may be increased, especially if n_b for case (2) is lowered.

It may now be more evident that adopting $|E| \approx |V_{dc}|$ at the base speed n_b is also good for generator operation in terms of energy conversion. This is not so in terms of excitation penalty. In general, an excitation penalty $\varepsilon = 0.3$ to 0.4 is considered acceptable, though higher values may lead to higher energy conversion in the machine. The $|E| = |V_{dc}|$ condition, if maintained for the entire speed range (70 to 140%, for wind generators), corresponds to the lowest generator peak current for given output power (energy per cycle) at maximum speed.

For wide constant speed range generating, it seems that choosing $V_{dc} = \text{const} = E$ at base speed is the solution. Designing the PWM converter at V_{dc} corresponding to base speed n_b pays off in terms of converter voltage rating.

9.6 PWM Converters for SRGs

As already alluded to in this chapter, each phase of SRG(M) is connected to a two-quadrant DC–DC converter (Figure 9.12).

The generator case is presented in Figure 9.12 for two main situations in terms of load — with and without backup battery. Only in the case with load backup battery may the SRG operate as a starter, supplied from the battery. In vehicular applications, this is the case for internal combustion engine (ICE) starting and driving assistance. The generator mode appears, during exclusive ICE driving, for optimum battery recharge and load supply and during regenerative vehicle braking. The low-voltage self-excitation battery is present and serves to initiate the raising of output voltage under no load (self-excitation) only when the load backup battery is absent.

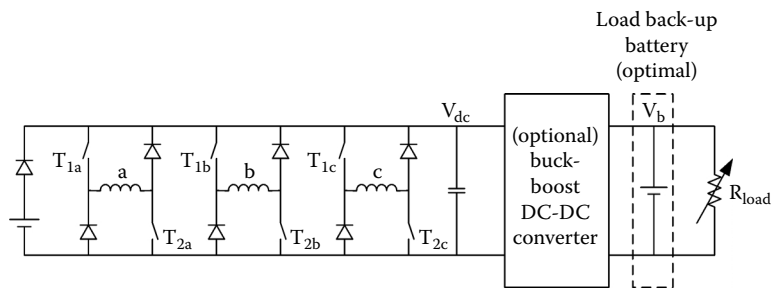


FIGURE 9.12 Three-phase switched reluctance generator (SRG) with asymmetrical converter, self-excitation, or load backup battery.

There are many other PWM converter configurations that use a smaller number of SCRs (m , or $m + 1$, instead of $2m$ pieces for the asymmetrical converter). A complete investigation of such converters is presented in Reference [2]. The conclusion is that all have advantages and disadvantages. At least for SRGs, where the generator control, especially, may encounter instabilities in DC output voltage control, due to inadvertent self-excitation, the asymmetrical converter (Figure 9.12) is considered a very practical solution. This converter provides for the freedom of soft (one SCR off, T_1) or hard (both SCRs off) de-energization of phases, to reduce output voltage pulsations. Also, to increase the generated power, an intermediary soft turning-off period (T_1 off), before the hard turning-off period (T_1, T_2 off) is feasible.

A DC–DC converter for voltage buck boost may be added between the load backup battery and the SRG side multiphase chopper (Figure 9.12). The battery voltage may be too small in some cases ($42 V_{dc}$ in a mild hybrid vehicle), so it pays off to design the SRG and the machine-side converter at a higher and constant voltage, which should be at least equal to the battery maximum voltage V_{bmax} .

As shown in Chapter 8, for the PM-assisted reluctance synchronous machine (RSM) starter/generator PWM converter, it seems to pay off to include an additional DC–DC converter for voltage boost (in motoring) and step down (for generating), both in terms of silicon costs and system losses.

To stabilize the output voltage, it may be feasible, for stand-alone DC loads, to supply the excitation power (phase energization) from a separate (excitation) battery (Figure 9.13). As the excitation power may amount to 30% (or more) of the output power, the excitation battery has to be strong. To avoid adsorbing large currents from the excitation battery, the diode D_{oe} may be used [9]. The presence of the starting battery allows for operation during load faults and for clearing them out rather quickly (Figure 9.14) [9]. The battery is also used for starting the prime mover of the SRG when the latter works as a starter for rather rare starts (as for on board aircraft).

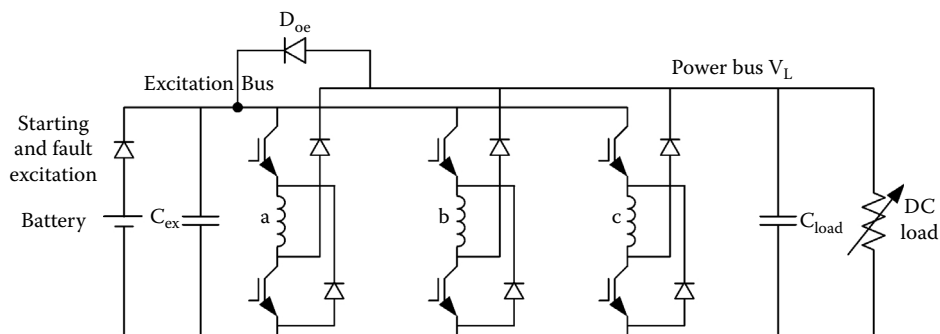


FIGURE 9.13 Switched reluctance generator (SRG) with separate excitation and power bus and fault clearing capability. (Adapted from A.V. Radun, C.A. Ferreira, and E. Richter, *IEEE Trans.*, IA-34, 5, 1998, pp. 1106–1109.)

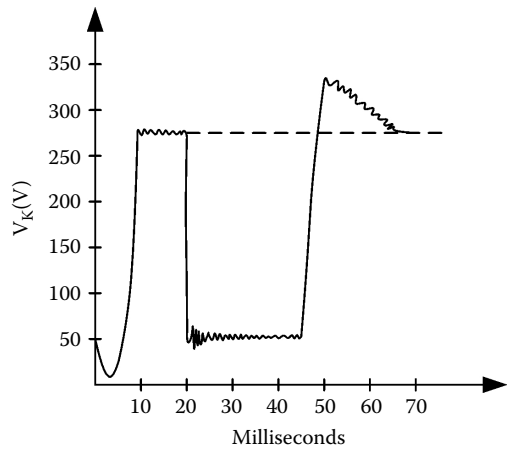


FIGURE 9.14 Voltage rise up, experiencing a fault (sharp load resistance decrease), clearing the fault, and voltage recovery.

Special means for the battery recharge are necessary with the scheme in [Figure 9.13](#), but at a low power rating. When one phase goes off, the SRG may provide continuous operation, at low load power, though, and with higher output voltage ripple as only the $m1$, $m2$, and $m3$ phases are sound. This is a special feature of an SRG, with phases that are weakly coupled magnetically, even for heavy magnetic saturation conditions.

The converter for the case of SRG with AC output should contain one more stage: a DC–AC PWM converter (inverter) ([Figure 9.15](#)). For such applications with speed variation not more than $\pm 25\%$ around rated speed, it may be practical to design the SRG to be capable of delivering constant (rated) V_{dc} — DC link voltage from minimum speed upward.

The AC load may be independent or it could be a weak (or strong) power grid. The control of the PWM inverter should be adapted to each of this kind of AC load. The capacitor C_{dc} may be designed to provide all the reactive power that the AC load is expected to absorb. More information on SRGs at the power grid is provided in Reference [10].

Note that it should be recognized that the machine-side converter is special for SRG in comparison with standard AC starter/alternators, but it contains about the same number of SCRs per phase (2 for four-phase machines). However, it avoids shoot-through short-circuits and is thus phase fault tolerant.

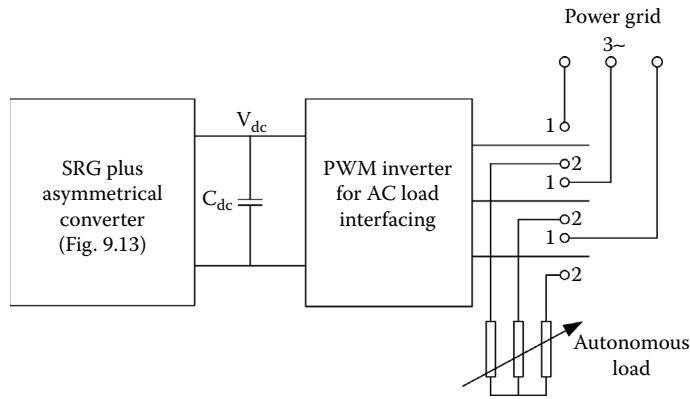


FIGURE 9.15 Switched reluctance generator (SRG) with additional alternating current (AC) load interface inverter.

Finally, the danger of emf that is too high at largest speed, existent in PM generators, is not present with SRG.

We treated here three- and four-phase SRGs, but one- or two-phase SRGs may also be built for low powers or superhigh speeds [11].

9.7 Control of SRG(M)s

There is a very rich literature on SRM drives control, with pertinent updates in Reference [2] and Reference [4, 12].

Though starter/generators experience longer generator than motor operation mode intervals, most attention has so far been placed on the motoring control.

We will distinguish two main high-grade control strategies for SRG(M)s:

- Feed-forward torque sharing control [2, 4]
- Direct torque control [14–16]

Current profiling with efficient torque sharing in four-phase SRG(M)s is also considered here as a feed-forward — indirect — torque control, though executed under a current profile control.

The control of SRGs may be classified by the controlled variable in the following:

- Speed or voltage control for generation mode
- Torque control for starter/generators

Speed control for maximum wind power extraction is typical for wind generators, but a torque (power) interior closed loop may be used for faster response.

Torque control may be used for generating with load backup battery, as the reference torque may be calculated from the required (accepted) current by the battery state of charge. The control system may be implemented with or without a motion sensor (encoder or resolver).

For mainly generator operation, even with infrequent starting (motoring), the rather coarse (if any) position feedback may be used. During, drive-assisting operation mode on EHVs, fast nonhesitant robust and precise torque response is required from zero speed. In this latter case, either a precise encoder is used to measure the rotor position, or an advanced position estimator that works from zero speed and has initial rotor position estimation capability is provided. As this kind of sensorless vector control was already demonstrated for PM-RSM starter/alternators ([Chapter 8](#)), such a standard behavior is expected from SRG(M), also.

In view of the above, we will present, in what follows, representative methods with feed-forward and with direct torque control. And then, we will deal separately with observers for motion-sensorless control.

9.7.1 Feed-Forward Torque Control of SRG(M) with Position Feedback

In automotive applications and, in general, for torque feed-forward control, the reference torque T_e^* is determined by the acceleration or braking pedal corroborated with speed and battery state and load information. It is the average reference torque, basically.

The magnetic saturation and the nonsinusoidal manner of phase inductance variation with rotor position make the relationship between torque and phase current highly nonlinear and speed dependent. Moreover, the phase current may be PWM controlled only below base speed, for motoring. To fully determine the torque produced by a phase not only the current i_i should be referenced (for PWM current control below base speed), but also the turn-on, θ_{on} , and commutation (hard commutation), θ_c , angles have to be referenced (changed).

Above base speed, when only the single-pulse current operation mode is available, for motoring, only the θ_{on}^* , θ_c^* angles have to be referenced for given T_e^* , battery voltage, and speed. The reference torque T_e^* vs. speed envelope has to be known off-line, to avoid control instabilities at high speeds for all battery voltages.

Feed-forward torque control is based on the off-line computation of θ_{on}^* , θ_c^* for given T_e^* , speed n , and battery voltage, based on the $\Psi(i, \theta_r)$ curves and the nonlinear circuit model of SRG(M), as described in the previous section.

To reduce the complexity of the various look-up tables required for real-time control implementation, only one phase torque is considered in three-phase SRGs. Torque ripple minimization is obtained with current profiling below a certain speed, calculated for minimum battery voltage [16]. When, at low speeds, PWM current control is used, the latter is applied only to one phase at a time.

The relationship between T_e^* , i_i^* , θ_{on}^* , θ_c^* for a large number of speed n and battery voltage V_{dc}^* levels, both for motoring and generating, are obtained offline from the machine nonlinear model via some analytical approximations.

In Reference [13], parabolic fitting curves are used:

$$\begin{aligned}\theta_{on}^*(T_e^*, n^*) &= \theta_z(n) + C_o(n)T_e^{m_o} \\ \theta_c^*(T_e^*, n^*) &= \theta_z(n) + C_c(n)T_e^{m_c} \\ i^* &= p_i(n)T_e + q_i(n)\sqrt{T_e}\end{aligned}\tag{9.56}$$

The parameters θ_z , C_o , C_c , m_o , m_c , p_i , and q_i are dependent on speed n and also on battery voltage V_{dc} .

A linear dependence of p_i and q_i on V_{dc} , may then be adopted [14]:

$$\begin{aligned}p_i &= k_p(n) + l_p(n)V_{dc} \\ q_i &= k_q(n) + l_q(n)V_{dc}\end{aligned}\tag{9.57}$$

Above base speed, the current regulation is no longer imposed, and only the angles θ_{on}^* , θ_c^* are imposed as dependent on speed.

It was proven [13] that the torque variation due to turn-on angle θ_{on} deviation is rather large, but it decreases when the torque increases.

Typical values of these off-line calculated control parameters for a 15 kW, 95 Nm, 12/8 (three-phase) SRG are shown in Figure 9.16 for motoring and in Figure 9.17 for generating, making use of the criterion of maximum torque for given current [13]. An extension of the torque/speed envelope is obtained when switching from this criterion to maximum torque/flux at high speeds [12]. The basic control scheme, shown in Figure 9.18, evidences the torque to current i_i , angles θ_{on}^* and θ_c^* , and the current regulators. Typical torque response at three speeds is shown in Figure 9.19 with current and torque vs. time for motoring and generating in Figure 9.20 and Figure 9.21 [14]. Good quality responses are visible in Figure 9.20 and Figure 9.21 for single-pulse current control (high speed).

At low speeds, below 100 rpm in Reference [16], current profiling may be used to further reduce torque pulsation and noise. Typical such current profiles with simulated torque are shown in Figure 9.22 [16] for a 24/16 three-phase 350 Nm FRG(M) at 200 V_{dc} and at a peak current of 200 A.

While the off-line computation effort of T_e^* , i_i^* , θ_{on}^* , θ_c^* for given n and V_{dc} is done once, at implementation, it is very challenging in terms of memory for online control with 50 μsec or so control decision cycles [16]. Rather good efficiency levels were demonstrated, however, by tests in Reference [16] (Figure 9.23a and Figure 9.23b).

To further reduce the DC current ripple that is felt by the battery, a soft commutation stage ($V_{dc} = 0$, one SCR only off) per phase, prior to the hard commutation stage, up to a certain speed, may be implemented [12, 16]. To avoid part of the tedious off-line computation of i_i^* , θ_{on}^* , θ_c^* for given T_e^* , n , V_{dc} various online estimation methods, such as fuzzy logic and ANN were proposed. A pertinent review of these control methods for motoring is given in Reference [12].

One step further in this direction is the concept of direct (close-loop) torque control of SRG(M).

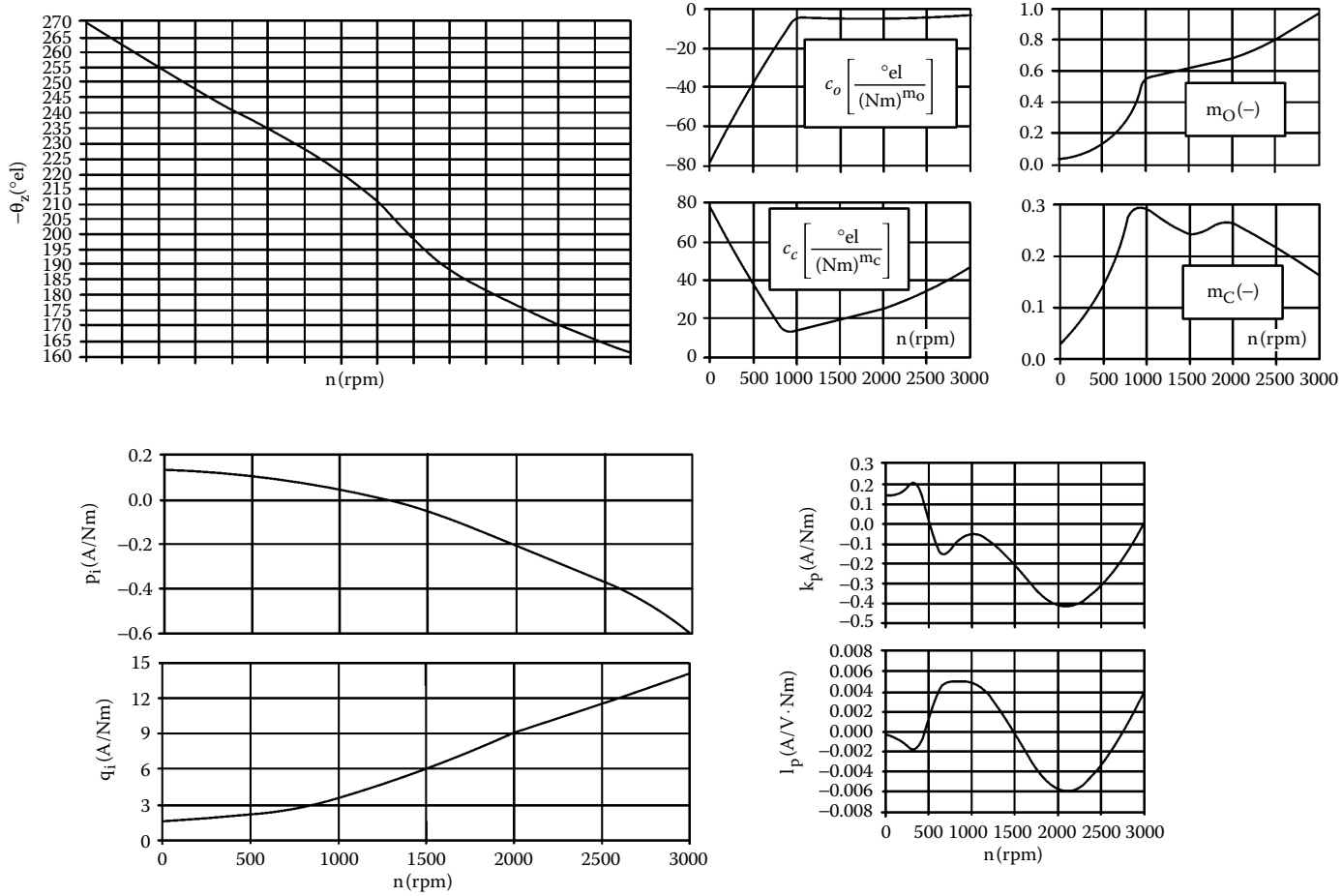


FIGURE 9.16 Precalculated control coefficients for motoring. (Adapted from H. Bausch, A. Grief, K. Kanelis, and A. Mickel, Torque control of battery-supplied switched reluctance drives for electrical vehicles, Record of ICEM-1998, pp. 229–234.)

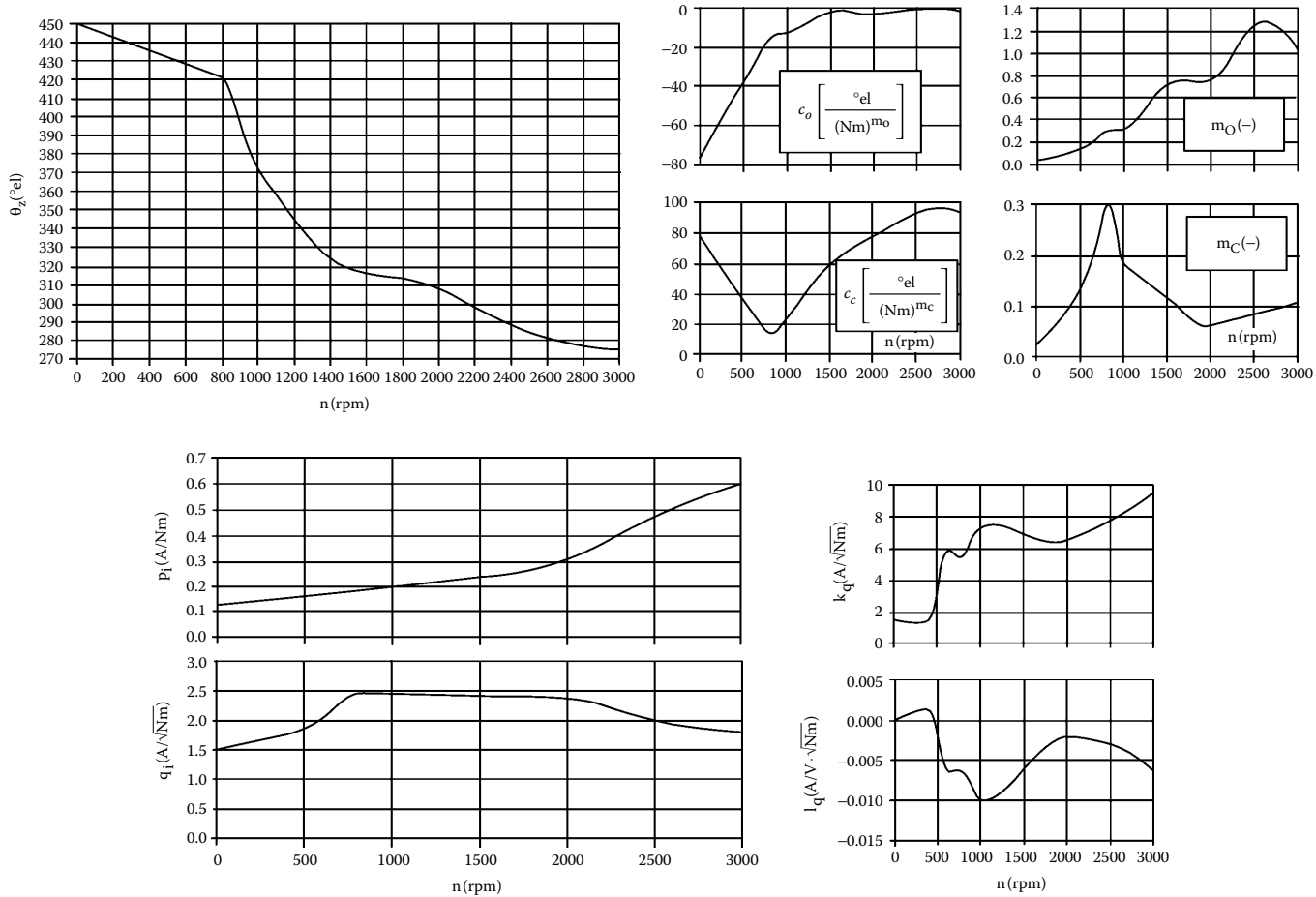


FIGURE 9.17 Precalculated control coefficients for generating. (Adapted from H. Bausch, A. Grief, K. Kanelis, and A. Mickel, Torque control of battery-supplied switched reluctance drives for electrical vehicles, Record of ICEM-1998, pp. 229-234.)

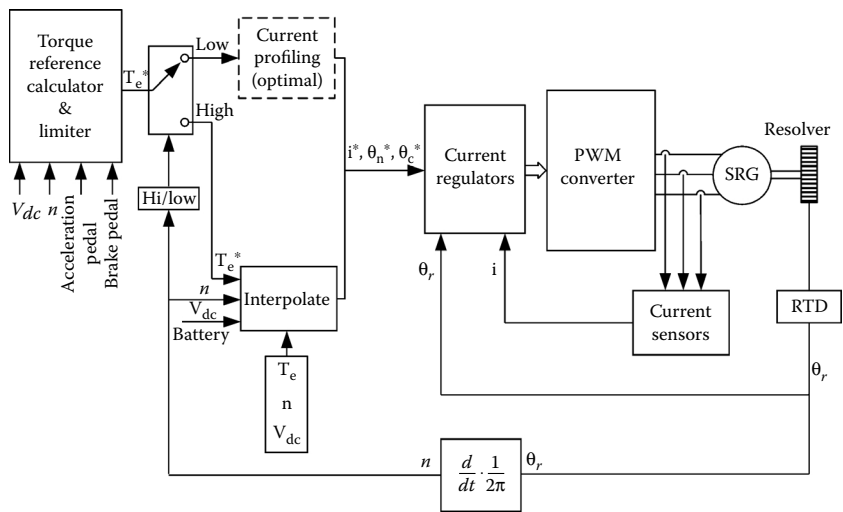


FIGURE 9.18 Feed-forward torque control of switched reluctance generator (SRG) (M). (Adapted from H. Bausch, A. Grief, K. Kanelis, and A. Mickel, Torque control of battery-supplied switched reluctance drives for electrical vehicles, Record of ICEM–1998, pp. 229–234.)

9.8 Direct Torque Control of SRG(M)

Direct torque control [12, 15, 17] requires online torque estimation. Torque estimation, in turn, requires phase flux estimation. The reference torque is the average torque. For three-phase SRG, the average torque per energy cycle is determined basically for one active phase, but with four-phase machines, both active phases have to be considered. To save online computation effort, the flux and average torque of a single phase is estimated. However, this limits the average torque control response quickness to fast reference torque variations. Estimating the flux and torque of all phases would eventually bring superior results in torque response quickness and quality, but with a markedly larger hardware and software effort.

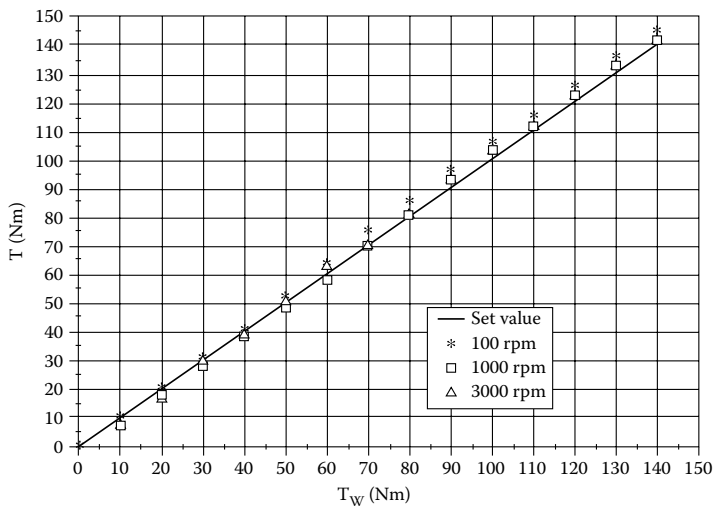


FIGURE 9.19 Torque response precision.

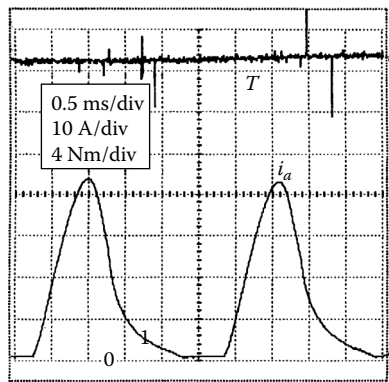


FIGURE 9.20 Current and torque for motoring at 1600 rpm and 32 Nm.

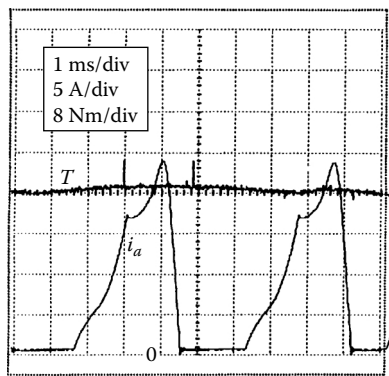


FIGURE 9.21 Current and torque for generating at 1600 rpm and 32 Nm.

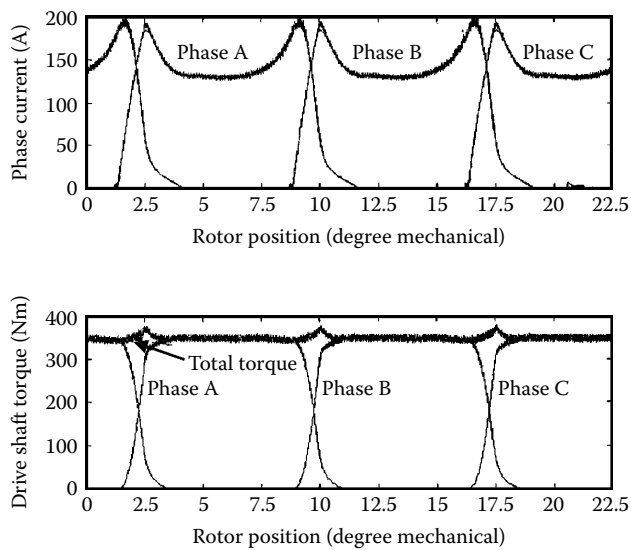


FIGURE 9.22 Simulated current profiles and torque.

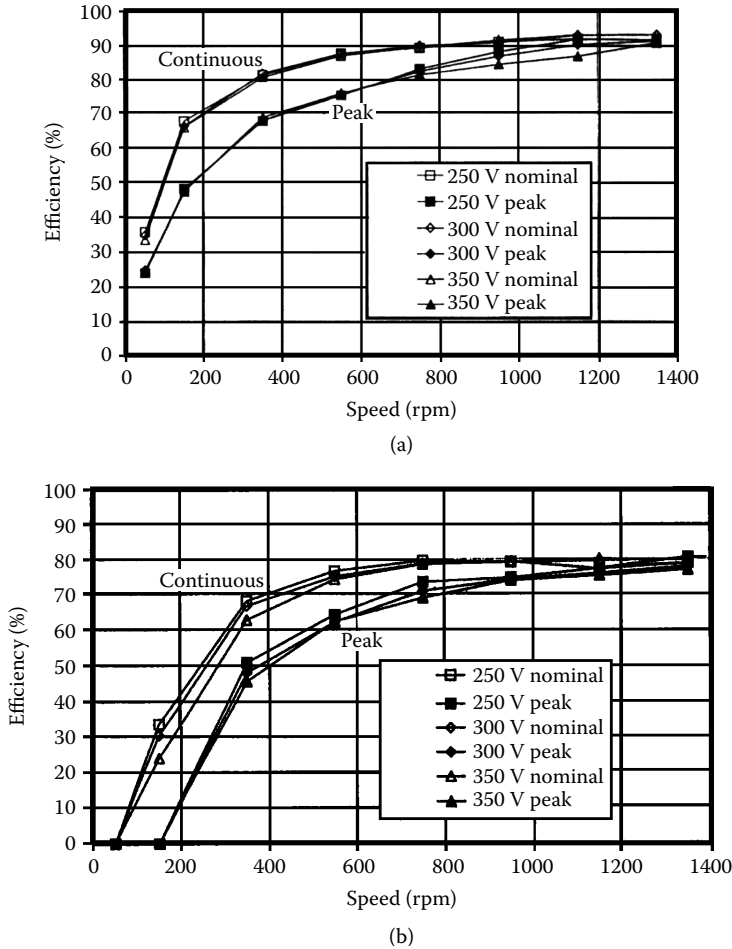


FIGURE 9.23 Efficiency vs. speed: (a) motoring and (b) generating.

Average torque T_{avr} estimation per energy cycle thus seems to be a practical solution. The problem is that torque estimation has to be sound, even at zero speed, in order to provide adequate torque control at zero speed. The presence of a refined position sensor is an asset in this enterprise, as precision in θ_{on}^* , θ_c^* control of less than 1° (mechanical) is required for 6/4 machines and less than 0.2° (mechanical) for 16/12 machines [17].

To calculate the average torque, the basic coenergy W_c formula from coenergy, via the mechanical energy per cycle W_{mec} is used:

$$(W_{mec})_{cycle} = \int_{\theta_{on}}^{\theta_{off}} \frac{\partial W_c}{\partial \theta_r} d\theta_r = \int_{\theta_{on}}^{\theta_{off}} \int_0^i \frac{\partial \Psi(\theta_r, i)}{\partial \theta} di d\theta_r \quad (9.58)$$

The energy required for phase magnetization is eliminated from Equation 9.58, because it is “recovered” during phase demagnetization.

As expected, W_{mec} contains the core loss, so it is not exactly the shaft torque. The average machine torque T_{avr} is

$$T_{avr} = \frac{m \cdot N_r \cdot W_{mec}}{2\pi} \quad (9.59)$$

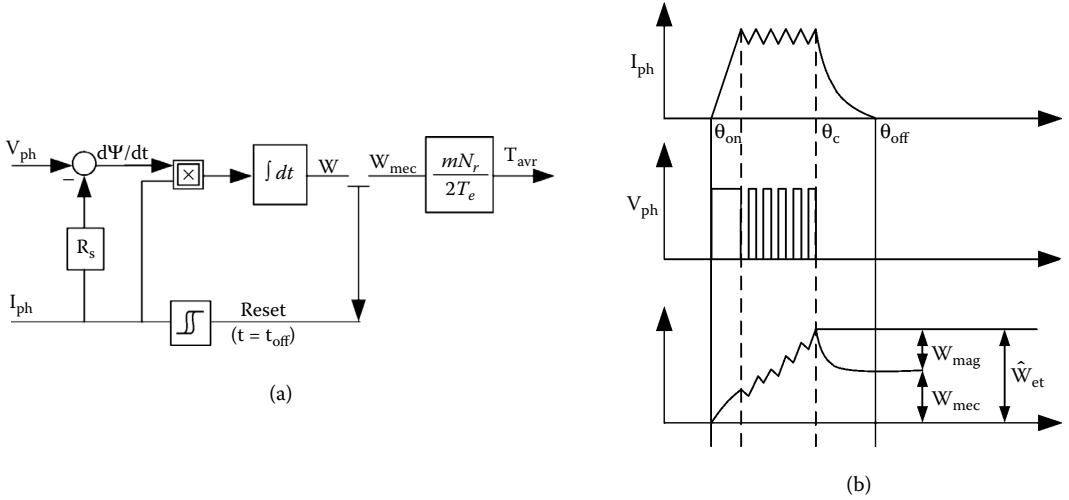


FIGURE 9.24 (a) Average torque estimator and (b) typical output at low speed (pulse-width modulator [PWM] current).

As the magnetization–demagnetization energies eliminate each other, W_{mec} may be calculated as follows:

$$W_{mec} = \oint_{cycle} \frac{d\Psi}{dt} i dt \quad (9.60)$$

The integral should be reset to zero when the phase current reaches zero, after each energy cycle.

But, the flux time derivative is straightforward:

$$\frac{d\Psi}{dt} = V_{ph} - R_s \cdot i \quad (9.61)$$

Simple as it may seem, but, when implemented, for low speeds, the phase voltage due to PWM is noisy, and temperature introduces notable errors. The situation may be ameliorated by some filtering of the phase voltage and by R_s adaptation.

The basic average torque observer T_{avr} is shown in Figure 9.24a and Figure 9.24b.

More advanced W_{mec} estimators may be introduced to facilitate better performance at low speeds. They may make use of the machine approximate current model (flux/position/current), $\Psi(\theta, i)$, even through linear approximations (9.33). Analog implementation of the average torque estimator has proven efficient [17].

The torque error ϵ_T may be used to provide the stator current i^* , turn-on θ_{on}^* and turn-off θ_{off}^* increments through a PI regulator. These increments may then be added to the off-line control parameters, as is done for feed-forward torque control (as discussed in the previous paragraph). A typical such direct torque control system is shown in Figure 9.25.

Good torque tracking control was claimed with this method in Reference [17], where, for simplicity, $\theta_{co} = \text{constant}$.

The off-line computation effort remains stern, while the direct average torque control adds more quickness and robustness to torque reference tracking, as many parameters vary in time. The battery voltage, V_{dc} , varies, for example, within 1 sec, from its maximum to its minimum during peak motoring torque application for in-town driving of an electric vehicle (Figure 9.26).

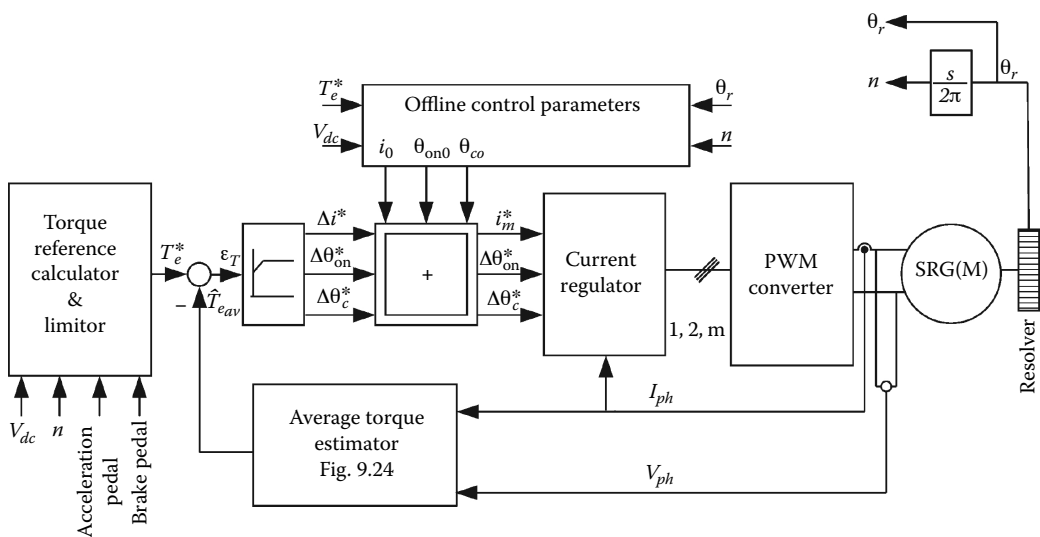


FIGURE 9.25 Direct average torque control of switched reluctance generator (SRG)(M).

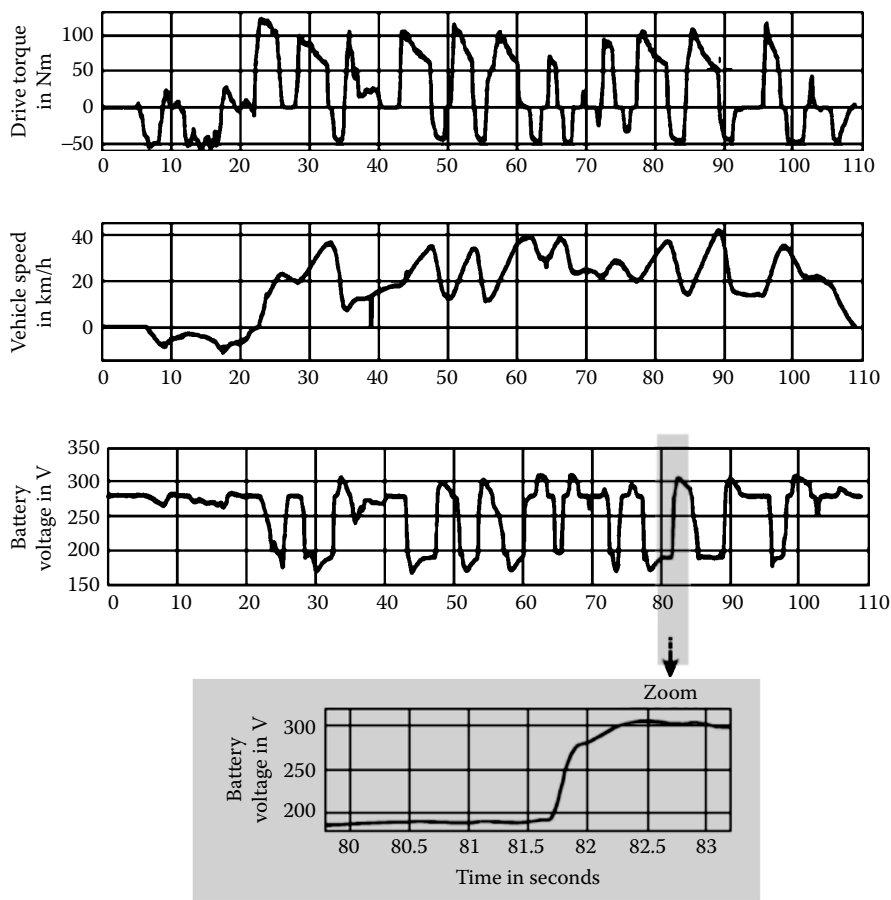


FIGURE 9.26 Torque, electrical vehicle (EV) speed, and battery voltage during standard town driving cycles.

9.9 Rotor Position and Speed Observers for Motion-Sensorless Control

In the control schemes so far presented, the rotor position and speed are provided by an encoder or a resolver with dedicated hardware for speed calculation.

As motion-sensorless control from zero speed, with signal injection for position estimation, was proven for PM-RSM starter/alternators ([Chapter 8](#)), the question is if the same can be done for SRG(M)s. There is a rich literature on motion-sensorless control of SRM without and with signal injection (in the passive phase in general). A pertinent review may be found in Reference [18].

9.9.1 Signal Injection for Standstill Position Estimation

To identify the initial rotor position for a nonhesitant start, a voltage pulse is applied to one phase, and the current is acquired up to a certain value i_k . The corresponding flux Ψ_k in the winding is as follows:

$$\Psi_k \approx V_{dc} \cdot t_k \quad (9.62)$$

where t_k is the time for full voltage application.

For a given Ψ_k, i_k pair, from the already stored $\Psi(\theta_r, i)$, which is eventually curve fitted, the initial rotor position is estimated. While this method is correct in principle, it does not provide rotor position estimation at low speed and up to the speed when emf methods may be used. At low speeds, including standstill, choosing an idle phase, a small AC voltage V_{sens} signal may be injected. Neglecting the emf, the equation of idle phase is as follows:

$$V_{sens} = R_s i_j + L_j(\theta_r) \frac{di_j}{dt} \quad (9.63)$$

With a sinusoidal voltage injection of frequency ω_s :

$$V_{sens} = V_m \sin \omega_s t \quad (9.64)$$

The current in the idle phase is as follows:

$$I_{sens} = \frac{V_m}{\sqrt{R_s^2 + \omega_s^2 L_j^2(\theta_r)}} \sin(\omega_s t - \gamma) \quad (9.65)$$

$$\gamma = \tan^{-1} \frac{\omega_s L_j(\theta_r)}{R_s} \quad (9.66)$$

Both the amplitude and the phase of I_{sens} contain position information.

The measured inductance, obtained through a modulation technique, may be used again, in corroboration with an appropriate $\Psi(\theta_r, i)$ family of curves, to determine the rotor position. The sensing voltage signal has to be “moved” from one idle to the next idle phase, and special hardware is needed to produce the sinusoidal sensing voltage and so forth.

More advanced rotor position and speed observers may be used for low and high speeds without signal injection. A sliding mode such observer, for example, may be of the following form [14]:

$$\begin{aligned}\hat{\theta}_r &= \hat{\omega} + K_\theta \operatorname{sgn}(e_f) \\ \hat{\omega}_r &= K_\omega \operatorname{sgn}(e_f)\end{aligned}\quad (9.67)$$

$\hat{\theta}_r$ and $\hat{\omega}_r$ are the observed rotor position and speed; k_θ, k_ω are so-called innovation gains, dependent on position and, respectively, on speed.

The error function of the observer e_f is as follows:

$$e_f = \sum_{j=1}^m \cos(\hat{\theta})(\hat{i}_j - i_j) \quad (9.68)$$

The estimated current is obtained after the phase flux is determined through integration:

$$\hat{\Psi}_j = \int (V_j - R_s i_j) dt \quad (9.69)$$

The current model based on known $\hat{\Psi}_j(\hat{\theta}_r, \hat{i}_j)$ retrieves, for already known $\hat{\theta}_r, \hat{\Psi}_j$, the estimated current $\hat{i}_j$. A combined voltage/current model may be used to estimate both $\hat{\Psi}_j$ and $\hat{i}_j$ at the same time, as done for AC machines (Chapter 7 and Chapter 8).

A motion-sensorless complete drive with interior torque close-loop control may be based on a parabolic relationship between flux Ψ and torque and a linear reduction of θ_{on} with speed and torque. The commutation angle θ_c varies from θ_c^i to θ_c^Ψ when the speed increases [14].

θ_c^i is the commutation (turn-off) angle for maximum torque per current; θ_c^Ψ is the commutation angle for maximum torque/flux. The latter criterion provides, for $\omega_r > \omega_b$, more available torque for given DC voltage, V_{dc} . Typical experimental results are shown in Figure 9.27a for responses at 95 rpm and in Figure 9.27b for those at 4890 rpm. [14]. Due to integrator drift, sensorless operation at very low speeds (below 95 rpm in Reference [14]) becomes difficult.

The combined voltage–current observer to estimate flux $\hat{\Psi}$ and current $\hat{i}$ may be used to reduce the speed for good position estimation, perhaps to a few (20) rpm [19, 20]. Still, for safe standstill torque control at zero and a few rpm, a signal injection rotor position estimator is required. Reference [21] suggests a rotating signal injection position estimation that works from zero speed and is similar to its counterpart for AC drives.

9.10 Output Voltage Control in SRG

Applications such as wind energy conversion or auxiliary power sources require constant DC voltage output for a limited speed range as mentioned before in this chapter.

A simplified control system for such applications relies on direct DC output voltage control with fixed or linearly decreasing with speed turn-on, θ_{on} , and commutation, θ_c , angles with rotor position feedback (Figure 9.28). An additional step-down chopper may be used to stabilize the DC output voltage control over a limited speed range.

The current regulator controls the level of machine energization through voltage PWM as the θ_{on} and θ_c angles are held constant. The self-excitation on no load is shown in Figure 9.29a to be slower than the start-up process on load with an external excitation source (a 12 V battery). Transient response to step reference DC output voltage shows rather fast response (Figure 9.29b) [22]. It is stable voltage response with zero steady-state error. The low voltage excitation source helps to self-excite the machine, especially when no remnant flux exists in the machine.

To produce more output, a limited soft turn-off stage — before hard turn-off at θ_c — may be introduced [23]. As a bonus, the noise level is also reduced.

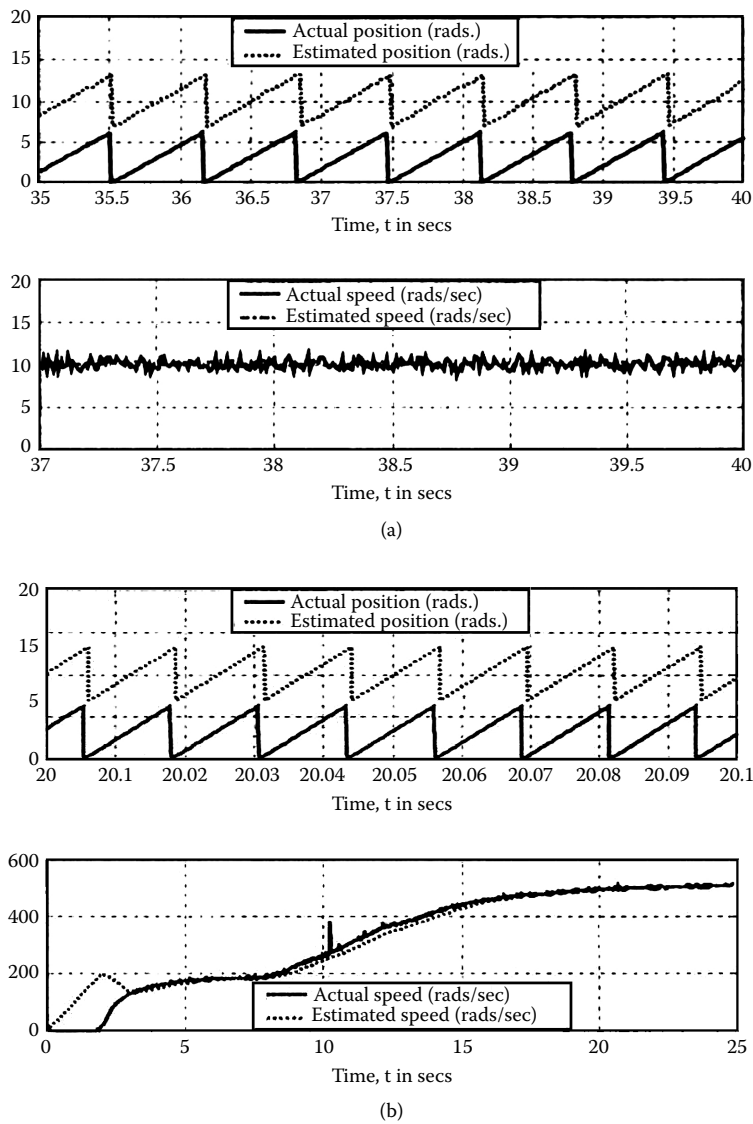


FIGURE 9.27 Position and speed observer responses at (a) 95 rpm and (b) 4890 rpm.

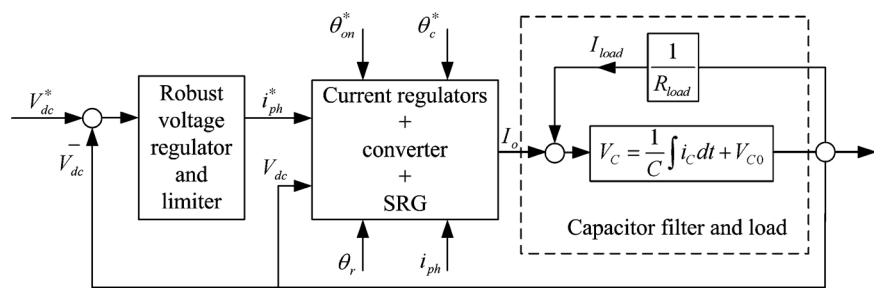


FIGURE 9.28 Direct current (DC) output voltage control of a switched reluctance generator (SRG).

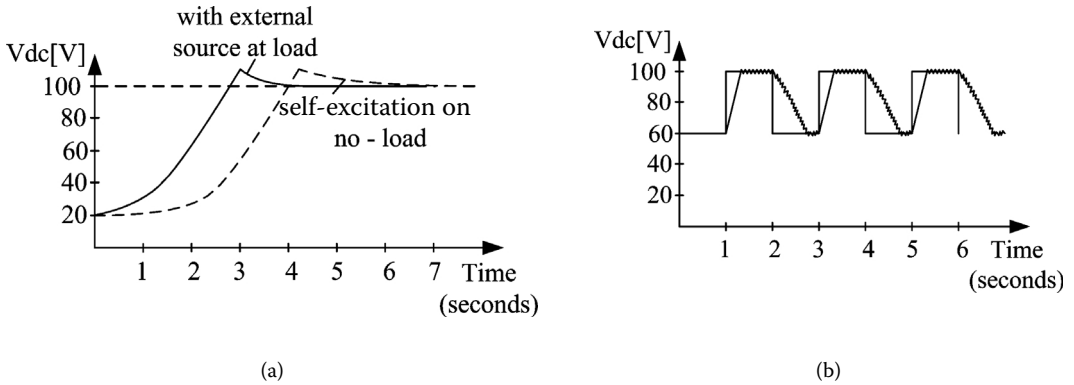


FIGURE 9.29 (a) The start-up and (b) voltage control.

9.11 Summary

- Switched reluctance machines are of double-saliency type, with coils, around all stator poles, supplied in sequence with current controlled voltage pulses in tact with rotor position. They have salient poles on the rotor and on the stator. Multiphase configurations are capable, with proper control, of rather smooth torque and self-starting capability from any initial position.
- Lacking windings or PMs on the rotor, SRG(M)s are rugged and suitable for hot environments (even up to 200°), and they are inexpensive.
- They work on the principle that the salient rotor poles are attracted to the energized phase (phases). As the rotor advances, one phase is turned off, while the next one is turned on at the adequate rotor position.
- SRMs are fully dependent on power electronics with rotor position-triggered control.
- Already, a handful of companies [3] commercialize SRM drives. As starter/generators, SRMs are vigorously proposed for automobiles, trucks, and aircraft applications.
- The machine configuration simplicity and ruggedness tend to be compensated for by the uniqueness, complexity, and costs of the PWM converter system.
- In the absence of magnetic saturation, the inductance of each phase has a three-stage variation with rotor position: rising, descending, and constant. The three stages correspond to $2\pi/N_r$ mechanical radians, where N_r is the number of rotor salient poles.
- In three-phase SRMs, only one phase has a nonconstant inductance vs. rotor position for either motoring or generating at any rotor position. There are two such active phases at all rotor positions in four-phase machines.
- The number of stator poles is $N_s = m \cdot 2K$, where m is the number of phases and $2K$ is the number of poles per phase.
- N_s and N_r are related by, in general, $N_s = N_r - 2K$.
- Three or four phases are typical in industrial SRM(G)s with $N_s/N_r = 6/4, 12/8, 24/16$ for three phases and $N_s/N_r = 8/6, 16/12, 32/24$ for four phases. To reduce the electromagnetic noise, an even number of pole pairs K per phase is chosen.
- The flux interaction between phases, even in the presence of magnetic saturation, is small. Hence, fault tolerance is rightfully claimed for SRG(M)s.
- Each phase is motoring, while its inductance has a rising slope and is generating along the descending slope. Unfortunately, each phase has to be fully fluxed and defluxed every energy stroke (cycle); that is, N_r times per revolution.
- The current polarity is not relevant to torque sign, because there is no interaction between phases and no PMs. This property leads to special PWM converters with unipolar current control.

- The family of flux/position/curves per phase $\Psi(\theta_r, i)$ is the most important performance identifier for SRM(G)s. While for aligned position the magnetization curve has to be highly saturated, the unaligned rotor position magnetic saturation curve is linear.
- The area between these two extreme curves represents the maximum mechanical energy W_{mec} converted into an energy cycle. For magnetization, the energy area is W_{mag} . The energy conversion ratio is $ECR = W_{mec} / (W_{mec} + W_{mag})$, which is 0.5 short of magnetic saturation and 0.65 to 0.75 for heavy saturation.
- Indicative for SRG(M) performance, in connection with the PWM converter, is the kW/peak kVA, which is only 10 to 15% lower than that for induction motor AC drives. However, 10 to 15% extra cost in the converter can weigh a lot in the system costs.
- The more common kW/rms/kVA ratio, or equivalent power factor, is not as important as the ratio between peak instantaneous current and peak RMS current that tends to be larger for SRG(M)s.
- The circuit model is based on independent phase voltage equations. Also, due to magnetic saturation, only a pseudo emf E may be defined, while the instantaneous torque has to be computed directly from $\partial W_c / \partial \theta_r$ (W_c is the coenergy per phase).
- The pseudo-emf E depends on rotor position and on current but is positive for motoring and negative for generating (Figure 9.6).
- The pulsed-supply of phases leads to iron losses both in the stator and rotor and makes the core loss model rather involved.
- The transient inductance $L_t = \partial \Psi / \partial i$ is, under saturation conditions, close to its unaligned value. This explains the acceptably quick phase fluxing during generating. Magnetic saturation also limits the maximum flux in the machine to keep the voltage rating within reasonable limits.
- Exponential, polynomial, fuzzy logic, ANN, or even piece-wise linear approximations were proposed to curve-fit the $\Psi(\theta_r, i)$ family of magnetization curves that is so necessary, both for SRG performance simulation and for control design implementation.
- The SRG design issues depend on the speed range for constant power and on output voltage.
- SRG(M) specifications are mainly expressed in terms of torque/speed envelopes for motoring and generating, for given DC voltage source (output) value. The maximum value of the current i_{peak} is also specified.
- After sizing the SRG(M), an optimization design is pursued based on maximum torque/current below base speed n_b (rpsec) and maximum torque per total losses, to maximum torque/flux, at large speeds (above n_b).
- The base speed n_b corresponds to the case when the pseudo-emf E equals a given fraction $\alpha_{pf} \approx 1$ of the DC voltage, and the machine produces the peak torque under the conditions of maximum torque per ampere.
- When a wide constant power speed range is required, with $\alpha_{pf} = 1$, the minimum speed n_{min} may be chosen smaller than n_b , but at the price of a lower number of turns/coil and higher peak current, and, consequently, higher converter costs (converter oversizing as for the AC starter/alternators [Chapter 7 and Chapter 8]).
- The current is chopped below n_b , and single current pulse operation takes place above n_b for motoring.
- Single-pulse generation is feasible for $n > n_b$, but PWM modulation may be supplied for all speeds to lower the torque during generation.
- The flat top current generating corresponds roughly to $E = V_{dc}$ and may be maintained for a range of speeds only if V_{dc} is allowed to increase with speed. In this case, for constant output (load) voltage V_{load} , an additional step-down DC-DC converter is used to interface the load.
- $E = V_{dc}$ corresponds to very good energy conversion in the machine and converter.
- Phase energizing for generating takes place along descending slope of phase inductance and, thus, is slow, and it takes its toll of P_{exc} in the power. When the phase is turned off (commutated), the free-wheeling diodes become active and deliver power P_{out} to the load. For high speeds, $E > V_{dc}$, while for low speeds, $E \leq V_{dc}$, if single pulse generating is performed with turn-on θ_{on} and

turn-off θ_c angles as controlled variables. The ratio $P_{exc} / P_{out} = \varepsilon$ is called the excitation penalty, which, for a well-designed machine, should lay in the interval of 0.3 to 0.5.

- The excitation penalty ε is not minimum when the energy conversion ratio in the machine is maximum ($E = V_{dc}$), but for $E > V_{dc}$.
- There are many unipolar current DC–DC multiphase converters suitable for SRG(M)s, but the asymmetrical one with two active power switches per phase is most used so far, especially for generators, where both hard- and soft-switching are required to increase power output and reduce noise and vibration.
- In essence, when SRGs supply a passive DC load, self-excitation is difficult, and a special low-voltage battery with a series diode is provided for initiation. For safe and stable output control, it is possible to use a separate excitation bus (or battery) and a power bus that might flow power back to the excitation bus, to lower the power rating of the excitation battery (Figure 9.13).
- Control of SRG(M)s for starter/generator applications amounts to feed-forward or direct (close-loop) average torque with or without position sensors.
- Average torque estimation proves to be rather straightforward, as the mechanical energy per cycle W_{mec} equals the resultant electrical energy per cycle $W_{mec} = \int \frac{d\Psi}{dt} i dt$ at the moment when the current in the respective phase becomes zero. This is so because the energy for fluxing and defluxing the phase cancel each other during a complete energy cycle. This is not so for continuous current control.
- The curve-fitted $\Psi(\theta, i)$ family is used off-line to calculate the optimum θ_{on} and θ_c angles and current, for given reference average torque, speed, and battery voltage and state of charge.
- So far, no practical control system that fully avoids off-line calculations was demonstrated for SRG(M)s in contrast to AC drives.
- Motion-sensorless control for starter/generators requires position estimation from zero speed. So, a signal injection method is mandatory for zero and low speeds, while an emf method is to be used when the speed increases. In References 21 and 24, such an efficient method of saliency detection through quasi-rotating voltage vectors injection and current response processing was proposed. It is a version to the one introduced earlier for AC drives.
- When only generator mode is required, simpler position estimators may be used, while direct DC output voltage control is performed.
- The control of SRG(M)s is more involved than for AC machines, but its capacity to perform over a wide constant power speed range, because phase defluxing is at hand, and machine ruggedness and low cost make the former a tough competitor to the AC starter/generator systems.

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